Rebuilt and Modified Altec 1567A: A Technical Report

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1. Introduction

This report describes the extensive re-build and modification of a 1960-era Altec 1567A microphone preamplifier/mixer. It is a form of “breakout modification,” in which individual input channels of a vintage preamp-mixer are given separate buffered outputs. This suits today’s multi-track studio ecosystem better than a dedicated monaural mixer. While such a modification need not eliminate the original mixer function, in this particular case it does (it can still be reconstituted externally, if desired). The old master channel is made completely independent (a fifth channel). A main goal here was to maximize options for—and control over—distortion and coloration by allowing the channels to be linked in different ways and in any order.

In this digital age, most audio engineers agree that well-maintained vintage outboard vacuum-tube gear lends body, warmth, and character to recorded tracks. However, opinions differ on the extent that the venerable old equipment should be modified. At one extreme, “purists” want to stay true to the vintage design. On the other hand, “pragmatists” welcome modifications that add versatility. This modified Altec 1567A is on the pragmatic end of the spectrum; obviously it is a “hot-rod,” not a simple restoration. It is an experimental, transformer-coupled, tube-based gain engine tailored to the adventurous audio engineer.

Solid-state line drivers buffer the outputs of the four transformer-coupled mic preamp channels. Relay-controlled pad, polarity, and impedance switches are added to their input circuits. Two of these channels are given variable attenuators (and an unbalanced input option) between their input transformer and triode stages; this allows experimenting with transformer saturation as a distortion effect. As a fifth independent channel, the old master channel is provided an unbalanced input and a few other modifications; it retains an all-tube signal path, tone controls, transformer-balanced output, and the VU meter. It can be an input channel (or DI) if desired.

The original Altec 1567A “vintage unit” was stripped down to the bare chassis before rebuilding with upgraded parts in critical cases, including all new ceramic tube and transformer sockets, and all new coupling and electrolytic capacitors. (Exceptions: original wiring and parts were left in the tone control and meter range networks.) The internal grounding scheme is upgraded to a hierarchical star-grounding rule, rather than multiple chassis tie points. To house most of the additional circuitry required by the modifications (including the solid-state output drivers, their power supply, input and output jacks, and additional knobs and switches), the vintage unit is permanently married to a four-rack-space black aluminum panel called the “auxiliary panel.”

Figure 1 shows front and rear views of the assembly, with some key parts labeled. It is held together with thick aluminum side panels and stiffened with two horizontal steel braces/handles in the rear. The entire 25-pound, seven-rack-space assembly is rack-mountable. Each channel’s I/O jacks and controls are aligned vertically between the two panels for an intuitive layout, except that most of channel 5’s controls and its VU meter occupy the top and right side of the vintage panel. Labels for controls and jacks have channel-specific colors for ease of use. Locating I/O jacks on the front of the auxiliary panel, rather than (less accessibly) on a rear panel, facilitates experimenting with patching the various channels together in different ways.

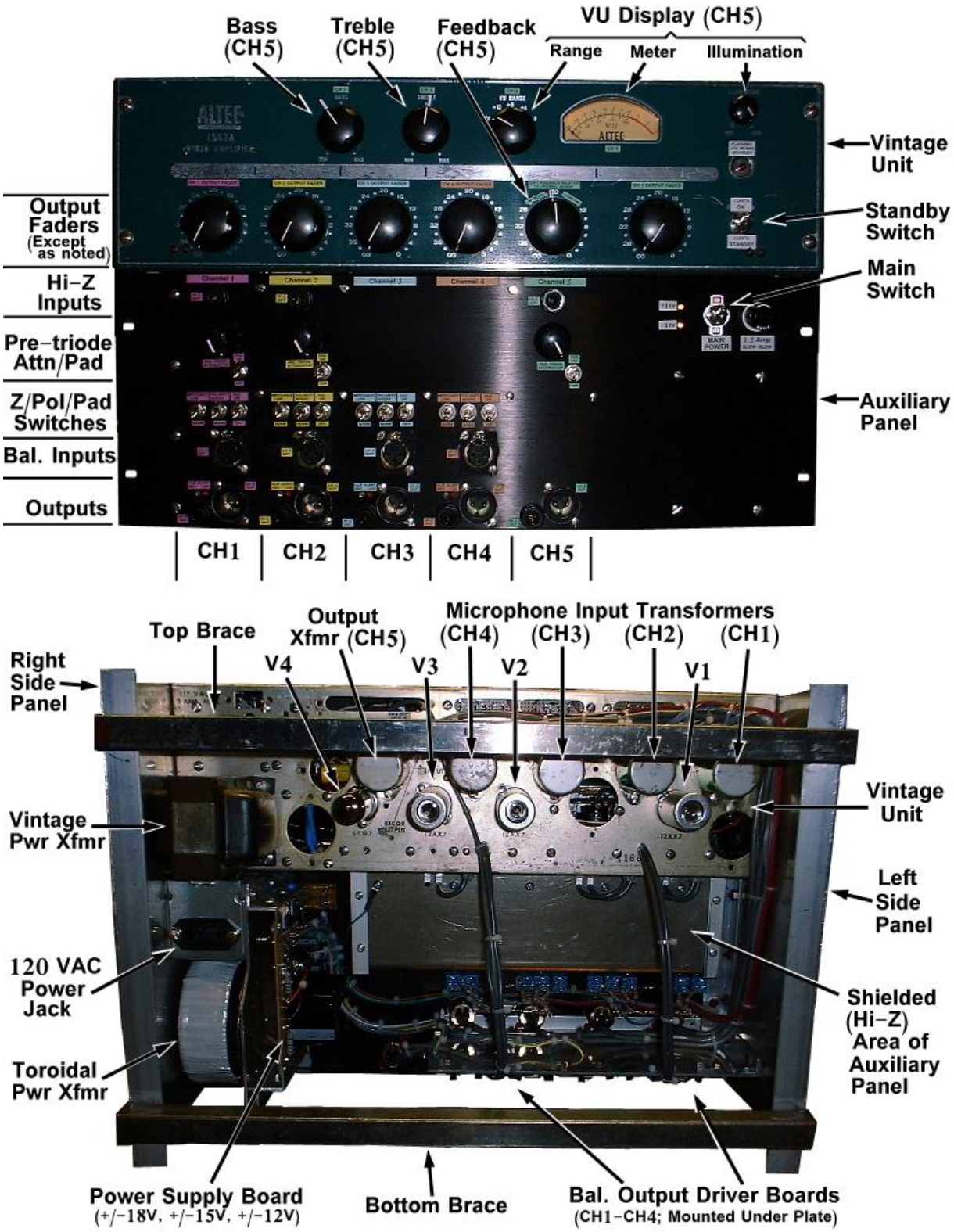


Figure 1. Front (upper image) and rear (lower image) views of modified Altec 1567A with some controls, assemblies, and parts labeled.

After briefly describing the original (stock) Altec 1567A, I will outline the modification's main features. After that, I will describe the modified unit in detail, and finally provide some observations on its performance and applications.

2. The Original Altec 1567A

2.1. General Description of Original Design. This initial description applies to all vintage Altec-Lansing model 1567A microphone preamp/mixers. Notes on the configuration and condition of the particular unit that was modified are in Section 2.2. For your convenience, a schematic diagram of the original (stock) Altec 1567A is reproduced in the Appendix at the end of this PDF document. Figure 2 is a block diagram of the audio signal path of a stock Altec 1567A configured as a microphone mixer. There are five input channels, a summing amplifier (mixer), and a master channel with three outputs.

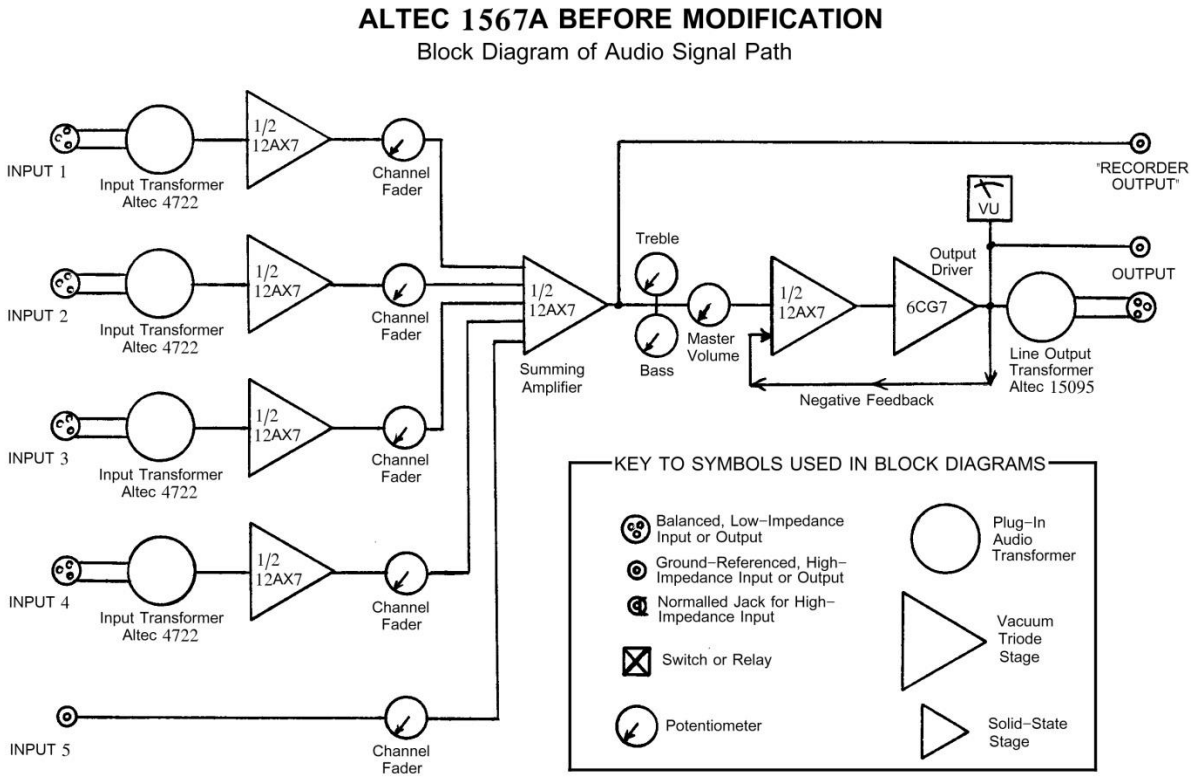


Figure 2. Block diagram of audio signal path in original (stock) Altec 1567A. The symbol key (inset at lower right) also applies to Figure 3's block diagram.

One input channel (Number 5) is passive and offers a high-impedance (high-Z), unbalanced input only. The others (1-4) each have a single-stage voltage amplifier using one triode of a 12AX7 vacuum tube. Altec-Lansing's way of making these active channels versatile was to provide octal sockets into which the following could be inserted: (1) a simple link between socket pins 5 and 7 for a high-Z unbalanced input, (2) a "phono equalizer assembly" for old-style

record players (channels three and four only), or (3) a model 4722 microphone transformer for a low-impedance (low-Z) balanced input. (Only the transformers are shown in Figure 2.) With its center-tapped primary winding, the 4722 offers either 150- or 38-Ω nominal input impedance.

Each input channel has a big knob labeled “MIX N,” where N = channel number, which is a channel fader in today’s jargon. These rotary fader pots feed a monaural mix bus (an assignment that cannot be changed without modification). Worth noting is an important difference between this typical 1960’s mixer head and today’s consoles: The vintage units lack gain controls analogous to the ones usually found at the top of modern channel strips. The gain of the vintage channels is always “set” at maximum, so useful fader settings are often in the lower (counter-clockwise) range of the rotary faders. Using modern consoles, engineers normally set channel gains so that faders linger near “0 dB” for the best signal headroom/noise compromise. In an Altec 1567A balanced input channel, the transformer and triode yield a combined voltage gain of about 60 dB. This is too much gain for some “hot” sources and no fader setting can give a clean signal (suggesting input pad options as a useful modification).

Fed by the mix bus, the summing amplifier uses one triode of a 12AX7 and is the first stage of the master channel. It drives a pre-tone-control “recorder output” and, via the tone control network (“treble” and “bass” knobs), the master volume control. The master volume control is followed in turn by the remaining 12AX7 triode, then a 6CG7 wired as a single triode serving as an output driver. Negative feedback between the latter two triode stages reduces the master channel’s gain, but helps reduce distortion, flattens frequency response, and lowers output impedance. The two main outputs are (1) an unbalanced output that does not depend on an output transformer being inserted and (2) a low-Z balanced output when a model 15095 line output transformer is plugged in. The original user manual (see link in Appendix) mentions that these two outputs can be used simultaneously. The 15095’s two secondary windings can be hooked in series or parallel to drive 600- or 150-Ω balanced lines, respectively. The VU meter is preceded by a 5-position rotary switch offering four sensitivities plus one “off” setting.

Overall, the Altec 1567A uses four twin-triode vacuum tubes: V1 and V2 are the 12AX7s for input channels 1-4, V3 is the master channel’s 12AX7, and V4 is the master channel’s 6CG7. All triodes are configured as common-cathode voltage amplifiers. Operating as high-gain, low-input-signal preamplifiers, the 12AX7s have shields and their heaters are powered with DC to minimize hum. The 6CG7 is on a separate AC-operated heater circuit shared with the VU meter’s two lamps. The power supply uses solid-state rectifier diodes.

2.2. Pre-Modification Condition of This Project’s Particular Altec 1567A. This section describes some specific issues with the unit noticed before its disassembly, and how some of these affected re-build and modification plans. The serial number of the vintage unit is 1188. Distinctive blemishes on the front panel are: a few deep scratches above and to the left and right of the power switch knob, and two small holes drilled under the “Altec” logo on either side of the words “MIXER AMPLIFIER.”

Old repairs to the unit were fairly easy to identify; none appeared recent (within the last many years). Green dye (factory quality-control?) was visible on unperturbed solder joints. Although the AC power cord was a replacement, there were multiple cracks in its outer sheath. (This was

not a concern because modification will provide a single standard IEC-style AC power jack behind the auxiliary panel.) Referring to part numbers in the original schematic (see Appendix), of the four chassis-mounted multi-section “can” electrolytic capacitors, C17A/B and C20A/B were replacements (attached by 4-40 screws and nuts instead of the factory rivets). Although C19A/B/C and C1A/B/C/D were original, C1B had been disconnected and replaced with a pair of 25-µF axial units in parallel. All other parts appeared original, including all signal coupling capacitors. Some modification or repair had been performed around the fader pots and mix bus on the front panel, but it was left wired according the original schematic.

According to the original manual, an optional assembly provides XLR input jacks and other connectors; this particular preamp was not so equipped, and screw-terminal strips were the only input-output (I/O) choice. Modification will eliminate these strips and provide I/O jacks on the auxiliary panel. The VU meter was also optional equipment, but this unit was equipped with a working one and a single working #44 lamp (half the number required). Lamp type is not specified in the schematic or manual, so I used two #47s, which run cooler than #44s.

The unit had four model 4722 input transformers and one model 15095 output transformer. The shield can on three of them seemed too loose, so I glued these to their bases with small dabs of epoxy cement at their crimp points. For the 4722’s, I marked the channels they came from (#1-4) upon extraction. All tested good for continuity and no shorts using an ohmmeter, but #3 had significantly lower-than-expected impedance on AC performance tests (more on this in Section 5.2). The two secondary windings of the 15095 output transformer did not have equal DC resistances (see Section 5.20). While later tests showed that audio sounded good using it, its function may be sub-optimal. As described in Section 4.10, I decided to add a rear-chassis-mounted switch for easy toggling between series and parallel secondary connections.

The unit came with three excellent Telefunken 12AX7 tubes, and one Raytheon 6CG7 tube. All of these tested good, and on a breadboard mimicking the vintage mic preamp circuit, each 12AX7 triode gave 35 dB gain (as loaded by a fader pot’s 250-KΩ resistance) at 1.0 KHz, and low noise. To coax maximum lifespan out of these particular 12AX7 gems, I decided it was worthwhile to design into the modification a soft-startup feature for them; this standby switch scheme is detailed in the vintage power supply circuit description (Section 4.3).

Before its tear-down, I performed a “smoke test” on the as-is unit, powering it up gradually through a “Variac” (variable line AC auto-transformer). The unit’s power transformer and other power supply components worked and nothing seemed to overheat. Most DC voltages were within 15 percent of those published on the schematic (Appendix); the plates of V4 were 20 percent low. I considered this a decent DC result for service-neglected gear of this age.

However, audio results were not “studio grade” by anyone’s standard. There was hum in the output, probably due to old and weak electrolytic filter capacitors. Of course, my re-build plan included replacing all of them, even upgrading to somewhat higher capacitances. (Cosmetically, the modified unit won’t look as “vintage” because I use axial electrolytics within the chassis, not the [now mostly obsolete] chassis-mounted “can” types. This also calls for laying out new terminal strips somewhat differently from the originals.) Also, modification will upgrade the grounding system to a more robust star-ground-based strategy. Not surprisingly, the as-is unit

had poor low-frequency response, probably due to the elderly signal coupling capacitors. My plan included replacing all of these with high-quality polypropylene film units.

While working, all of the original tube and transformer sockets were old, tired, and dirty. My preference when rebuilding old gear is that sockets should be replaced with new ones for highest reliability. I decided to use ones with ceramic insulators for the best long-term stable performance. Happily, all of the potentiometers operated without scratchy spots (once exercised a little), so none of them required replacement.

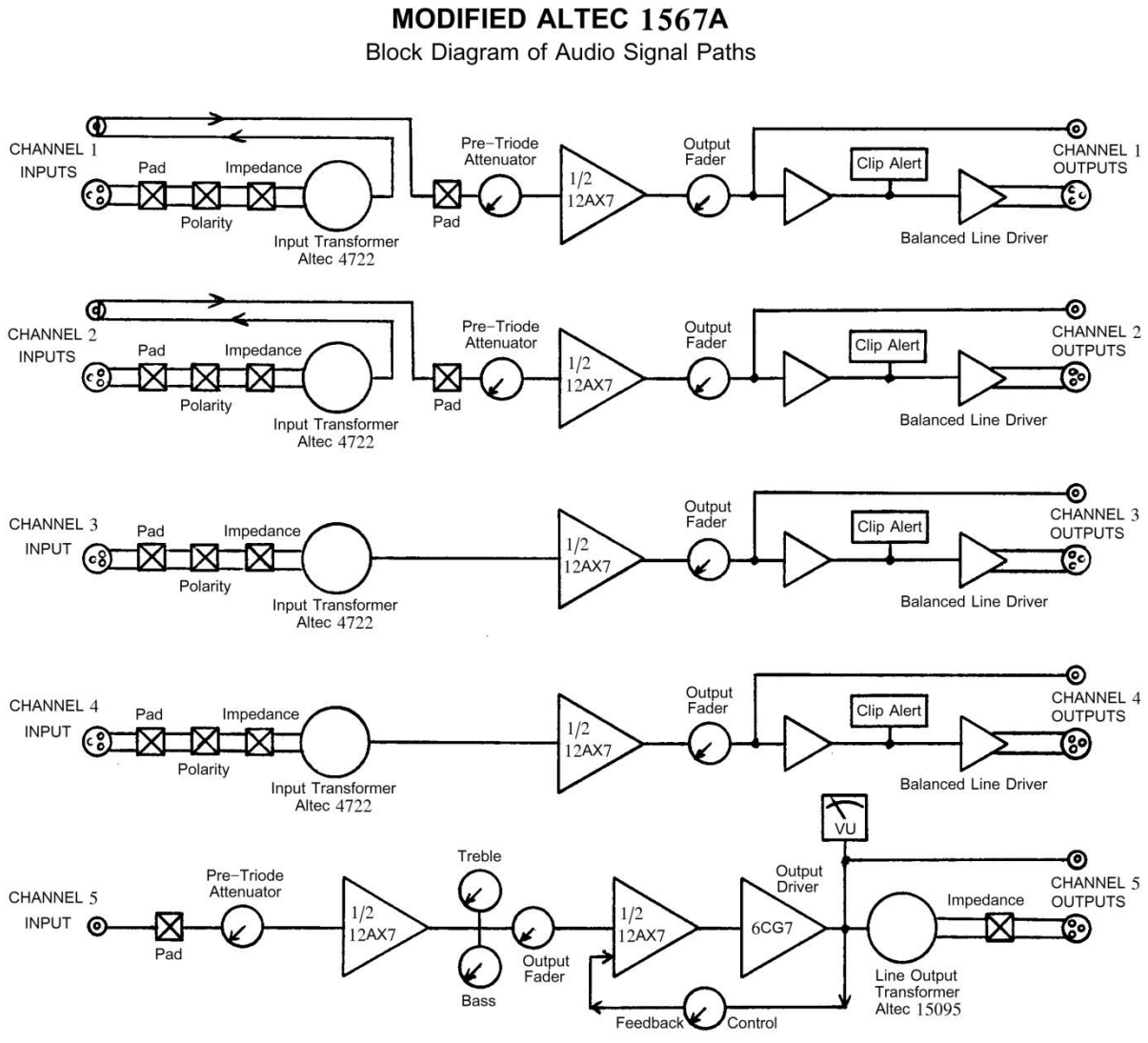


Figure 3. Block diagram of audio signal paths in modified Altec 1567A. See Figure 2's inset for key to the symbols used.

3. Overview of Modified Unit

3.1. General Architecture. Figure 3 is a block diagram of the modified Altec 1567A's signal paths. Compared to the original unit (Figure 2), notice that the modification trades the original mixer function for channel independence. Channels one through four (CH1-CH4) are based on the original unit's four microphone preamps. Channel five (CH5) is derived from the former master channel. Each channel has a balanced low-impedance (low-Z) output at a male XLR connector and an unbalanced high-impedance (high-Z) output at a ¼-inch female jack.

3.2. Channels 1 Through 4. These channels share a common basic layout, except that CH1 and CH2 have additional circuitry in the high-Z link between the input transformer and triode stage (see next paragraph). All transformer-coupled low-Z balanced input circuits have pad, polarity and Z-select switches. The normal setting for each toggle switch is the "down" position. These switches operate relays installed in the vintage chassis near the input transformers, so sealed relays with gold contacts handle the signals for enhanced reliability. The balanced outputs of CH1-CH4 are driven by solid-state buffers linked to the wiper of each channel's fader pot (more on this below). The direct wiper signals are available at unbalanced high-Z outputs.

3.3. Special High-Z Circuitry in Channels 1 and 2. Absent in CH3 and CH4, the additional high-Z circuitry in CH1 and CH2 is built in a shielded enclosure of the auxiliary panel; internal partitions help limit crosstalk between channels. Important performance limitations of this build-out are given in Section 5.13. Trade-offs aside, it adds the following features to CH1 and CH2:

> **High-Z Unbalanced Input.** This ¼-inch jack bypasses the input transformer to patch unbalanced signals directly to the triode grid circuit. At 1 KHz, input impedance is 840 KΩ when the pre-triode attenuator is full clockwise and the high-Z pad is off. This is a "normalled" jack, meaning it normally links the input transformer secondary to the triode. Inserting a ¼-inch plug there automatically opens that connection and replaces it with the ground-referenced signal at the plug's tip.
>
> **Pre-Triode Attenuator Pot.** The main practical application of this pot (and the pad switch described next) is to reduce a signal's amplitude when driving the input transformer to saturation for distortion effects. Without attenuation, the signal would be much too high to isolate the effect of transformer distortion, due to simultaneous triode super-saturation. The normal attenuator setting is full clockwise, but when experimenting with transformer saturation, this knob must be nearly fully counter-clockwise.
>
> **High-Z Pad.** The normal position of this toggle switch is down. The up position engages a -20 dB pad that expands the useful counter-clockwise range of the attenuator knob. It may be useful when applying an exceptionally hot signal to the balanced input.

3.4. Channel 5. This channel has a high-Z input with a pad switch and attenuator pot built into the shielded high-Z enclosure of the auxiliary panel. Otherwise it is similar to the old master channel (retaining its all-tube signal path and the VU meter), except negative feedback is made variable using the original "MIX 5" fader pot (for the old passive input channel, eliminated upon

modification). Decreasing feedback (clockwise) yields more gain and harmonic distortion. CH5's transformer-coupled low-Z balanced output can be set for nominal 600- or 150-ohms (series or parallel secondary connection, respectively) using a toggle switch added to the rear of the vintage unit.

3.5. A Word About Solid-State Outputs. At least with respect to their balanced outputs, CH1-CH4 are "hybrid," meaning they have both tube and solid-state circuitry. The outputs of these channels each use a high-quality op-amp followed by a state-of-the-art THAT1646 balanced line driver chip. As a byproduct of the way the latter chip creates a transformer-like low-Z differential output, they contribute nearly 6 dB voltage gain (see Section 5.9). "Purist" tube audio philosophers may reflexively consider this an anathema, so let me digress briefly to defend hybrid architecture in this case:

No practical recording studio in the 21st century has an all-tube (or even all-analog) overall signal path. Tube stages are used for—let's face it—effect (pleasing non-linearity). At their best, solid-state stages are low-noise and very linear; in short, they are neutral. They are also inexpensive to implement. Since vintage non-linear stages need at some point to be interfaced with modern solid-state/digital gear anyway, why not use sonically neutral solid-state stages within modified vintage gear if this adds versatility and connectivity? There are few non-emotional reasons why not. However, attention is required to the abrupt limit of the linear range of solid-state output buffers (i.e., hard clipping)—a decidedly non-vintage form of distortion. This modified Altec 1567A's "clip alert" indicator LEDs address this, as noted next.

3.6. Clip Alert LEDs for CH1 Through CH4. When saturated, the distortion mode of channel 1-4's output drivers is hard clipping. The "clip alert" feature is designed to help users tell if peak transients are exceeding clip thresholds, without monitoring with an oscilloscope. Trigger thresholds for the red "alert" LEDs are adjustable using 15-turn trim-pots accessed through the auxiliary panel. As presently adjusted, the alert LEDs light up exactly at the clip threshold only when driving floating 600-ohm loads. This occurs at a very high RMS output level of 25.5 dBV (53 volts peak-to-peak). In normal operation, it should only be a concern when driving another channel to transformer saturation. However, dependency of accurate clip indication on the load is an issue users must bear in mind; see Sections 4.9, 5.9, and 5.10.

3.7. Input/Output Polarity. For balanced inputs and outputs, all XLR jacks are wired according to the modern standard of pin 1 = ground, pin 2 = "+" or "hot", and pin 3 = "-" or "cold". Available on CH1, CH2, and CH5, the unbalanced inputs (1/4-inch jacks) have the same polarity as XLR pins 2 ("hot"). However, all five unbalanced outputs (1/4-inch jacks) match the polarity of XLR pins 3 (i.e., inverted with respect to the other inputs and outputs). The rationale is given in Sections 4.8 and 4.10.

3.8. Mechanical Layout and Power Supply. To wrap-up this overview, I will mention some infrastructural features that are not directly in the signal paths, and hence are not block-diagrammed in Figure 3. The rebuilt vintage unit and the auxiliary panel of the modified Altec 1567A (see Figure 1) are permanently married, both electronically and mechanically. It was logical to locate the auxiliary panel's power supply on the far right, and to put a single main power switch for the entire assembly there. This keeps the line AC circuitry as physically

compact as possible. It also eliminates the need for the original vintage unit's power switch; the modification places a standby switch for the 12AX7s at that physical location. A single fuse protecting the entire system is located on the auxiliary panel; the position of the original fuse is occupied by a flashing "standby" red LED indicator on the modified vintage panel.

4. Schematics and Circuit Descriptions of Modified Unit

4.1. Note About Schematic Diagrams. In this section, the design of the modified Altec 1567A is presented as a set of seven schematic diagrams (three for the power supplies, three for CH1-CH4, and one for CH5). You may need to zoom your PDF viewer to read the detail in these drawings. When comparing them to the schematic of the stock vintage unit (Appendix), please note that numbered components (e.g., SW2, R3, etc.) are not meant to correspond between the stock and modified units. Also, sequential numbering is reset for each different schematic of the modified unit. Therefore, text references to component numbers apply specifically just to the schematic being described. The only exception is the vacuum tube designations (V1-V4), which refer to the same parts in all schematics, both within this section and in the Appendix.

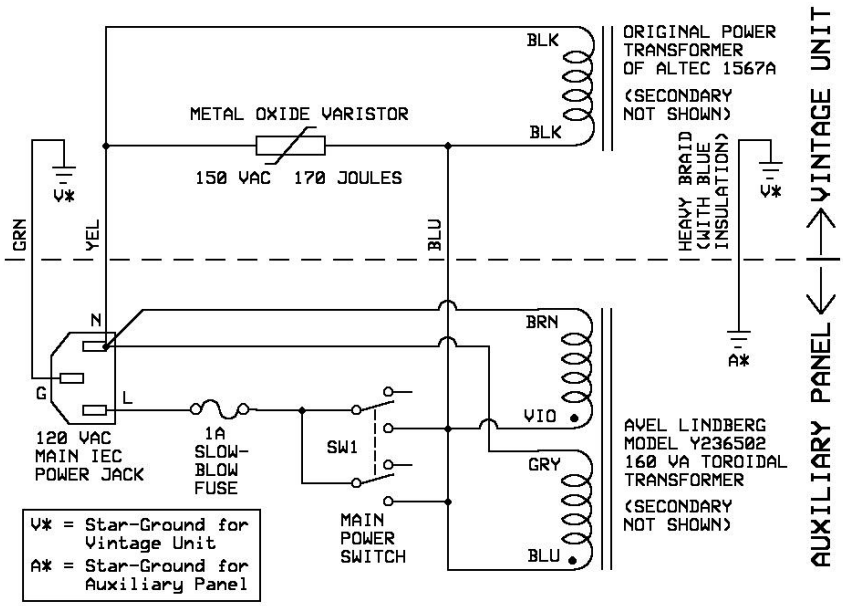


Figure 4. Schematic diagram of the modified Altec 1567A's line AC connections to the power transformer primary windings (secondary circuits are in Figures 5 and 6). The ground link between the vintage unit and auxiliary panel is also shown.

4.2. Power Supply: Line AC and Power Transformer Primaries. Figure 4 shows how the modified Altec 1567A's single 120 VAC power jack feeds the primary windings of the vintage unit's power transformer and the auxiliary panel's toroidal power transformer. The power jack, fuse holder, and main power switch (a DPDT toggle wired in parallel making an SPDT with higher reliability) share a small enclosure (AC box) behind the upper right corner of the auxiliary panel. This is near the vintage unit's power transformer, so only short wires are necessary to

reach the latter's primary via a hole punched in the vintage chassis directly over the AC box. The toroidal power transformer is mounted directly below the AC box, so its leads are also as short as possible; its primary windings are hooked in parallel for 120-VAC use only.

A metal oxide varistor (MOV) serving as a transient surge suppressor lends some overvoltage protection, mostly for the regulators in the solid-state power supply, but the old vintage power transformer may also benefit. The MOV is located in the vintage unit for ease of replacement should a large surge cause it to short out or explode. The fuse is a 1-A "slow-blow" type. Once warmed-up, the overall modified unit dissipates about 35 watts, which is 0.29 A at 120 V_{RMS}; the vintage unit accounts for about 0.16 A, and the auxiliary panel 0.13 A. But the inrush current upon power-up briefly exceeds 1 A, which a standard fast-acting 1-A fuse (as used in original vintage unit) doesn't always withstand. This surge is caused by filter capacitors charging, and also the considerable magnetic inertia of the big 160 V-A toroidal transformer.

Also shown in Figure 4 is the link between the star-grounds for the vintage unit and the auxiliary panel. It's included in this schematic because the low-resistance, heavy braided conductor making this connection passes through the same hole in the vintage chassis as the line AC connection. The star-ground terminal for the vintage unit is a short heavy-gauge solid copper wire anchored at its ends with two of the bolts mounting the vintage power transformer to the chassis (see Figure 9 for photo inside vintage unit; also see Section 4.4). The star-ground terminal for the auxiliary panel is a piece of thick brass foil mounted on one edge of that panel's power supply board (see Figure 7 for photo of this board).

The original schematic of the vintage power transformer (Appendix) shows an electrostatic shield (between the primary and secondary windings) that is hooked to ground; I noted that it is not terminated with a separate wire as in many isolation transformers. It's probably hooked internally to the transformer's housing/mounting base.

4.3. Power Supply: Vintage Unit. The schematic for the stock Altec 1567A (Appendix) shows the original power supply integrated with the overall unit. For the modified unit, I have drawn the corresponding power supply separately as Figure 5 (including the actual loads in the case of tube heaters and pilot lamps), which keeps the other schematics signal-path-oriented for less clutter. The vintage power transformer has three secondary windings: one for the high-voltage plate circuits ("B+"), one for the 12AX7 heaters (V1-V3), and one for the 6CG7 heater (V4) plus the two incandescent lamps for VU meter illumination. While it's logical to describe them separately, note that the standby switch (SW1) simultaneously affects both the heater circuit and the B+ supplies for V1-V3. Also note that the original rectifier diodes are replaced with modern units (D1-D4) for better performance and reliability.

The high-voltage supply uses "voltage doubling" rectification, since the bridge driven by the high-voltage secondary includes C1 and C2. Compared with the original, these capacitors have larger values (100 µF versus 60 µF); this additional capacity makes the bridge's output 364 V versus the original's 340 V (all voltages measured under normal operating load). To compensate, the first filter/voltage-divider network resistor (R1) is 4.7 K (versus the original 2.2 K) so that B+ for V4, and all B+ voltages down the line, matches the original design. In addition to R1's greater resistance, the modification also uses somewhat higher-value filter capacitors

throughout the network (and indeed an additional R-C stage; see below), than does the original power supply. This results in less ripple (hum) in the B+ lines compared to the stock unit.

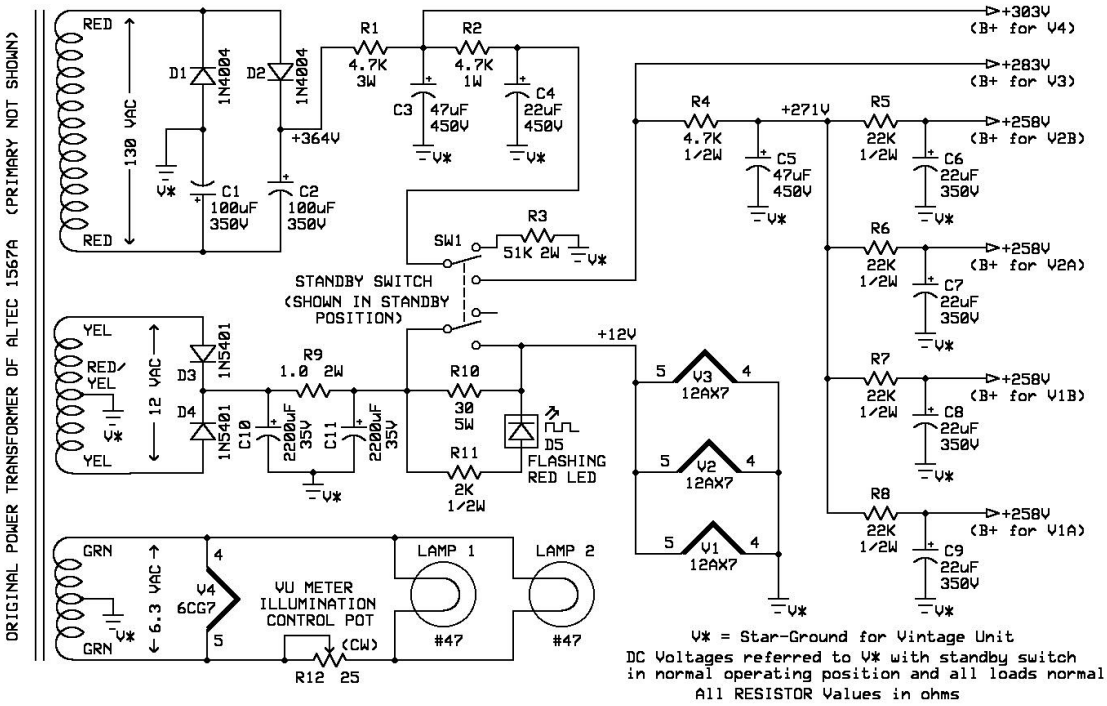


Figure 5. Schematic diagram of power supply in the modified Altec 1567A's vintage unit. See Figure 4 for the transformer's primary circuit.

After R1, the rest of the high-voltage filter/voltage-divider network serves the 12AX7s (V1-V3) via one-half of DPDT standby switch SW1. In standby mode (toggle handle in "down" position), high voltage to the six 12AX7 triode circuits is switched off, and R3 serves as a dummy load equivalent to these circuits. Standby mode is intended as a soft-start feature for the 12AX7s; while these tubes are warming up at reduced heater current (see below), keeping their high voltage turned off may help minimize "cathode stripping" and extend their lifespans.

In normal operating mode (standby toggle handle "up"), the DC characteristics of the B+ supply network for V1-V3 is equivalent to the original design; however, the four triode circuits of V1 and V2 (i.e., channels 1-4) are decoupled through individual R-C filters. In other words, R40 and C19A in the original schematic (see Appendix) was the single R-C stage common to all four triodes of V1 and V2; the modification (Figure 5) replaces this with series stage R4/C5 followed by parallel stages R5/C6, R6/C7, R7/C8, and R8/C9. These parallel stages respectively decouple V2B, V2A, V1B, and V1A. It is a more robust and stable approach, and extends cross-talk suppression among CH1-CH4 to higher frequencies than the original design provided.

The DC power supply for the 12AX7 heaters is like the original except for higher value filter capacitors C10 and C11 (2200 µF versus the original 1000 µF, for cleaner DC), and the addition of the standby switch network. In standby mode, the half of SW1 that is wired across R10 is

open. This resistor is in series with the heaters for V1-V3, so it limits the current through them (by about one-half), including the current surge on cold start-up. Therefore, powering up the modified preamp in standby mode should make 12AX7 heater failures less likely than in the original design. Via its current-limiting resistor R11, flashing red LED device D5 is powered by the voltage drop across R10 to advise users the 12AX7s are on standby. About 30 seconds after power-up in standby mode, SW1 can be switched to normal operating mode (standby toggle “up”), shorting R10 to extinguish D5 and complete power-up of the 12AX7s. Powering the unit down should use the reverse procedure (i.e., going to standby mode before main power off) so the standby switch is set to the recommended position before the next use.

The final circuit of the modified vintage unit’s power supply is for the AC-powered heater of V4 and the two meter-illumination lamps. It is unchanged from the original design. A mid-range setting of illumination pot R12 is recommended for maximum lamp lifespan.

4.4. Grounding Rule for the Modified Vintage Unit. As was routine in the Altec 1567A’s era, the chassis served as the ground-distribution network for most of the common (nominal zero-volt) nodes in the power supply and in the audio circuits. The vintage chassis-mounted multi-section electrolytic filter and cathode-bypass capacitors were equipped for this grounding method: twist-lock tabs securing them to their metal mounting flanges doubled as their common negative terminals, and provided convenient lugs for nearby circuits’ other ground connections. (Note in original schematic how, of the two 60-µF capacitors in the high-voltage supply’s bridge network, only C17B was in a multi-section unit, because it’s negative terminal was grounded; its partner C21 had to be an discrete axial-lead device since it floated above ground by about +170 V.) While this grounding strategy obviously works and simplifies manufacture, theoretically it is not ideal; and practically, mechanical connections to the chassis can corrode or loosen, causing long-term reliability concerns.

In contrast, an ideal star-ground is a single point on the chassis to which all ground connections are made. It is the one node whose potential is precisely zero volts by definition, and all ground connections are referred to it; hence the grounding network is shaped like a “star.” Granted, there is resistance in each wire to this node, causing local ground-potential errors, but these can’t interact with each other or accumulate as they might in a “mesh”- or “chain”-shaped, or even a chassis-based grounding network. An ideal star-ground scheme makes internal ground loops (which can couple hum or crosstalk into the audio path) impossible.

Having said all that, I hasten to state that the modified vintage unit described here does not strictly adhere to an ideal star-grounding rule, even though I use the term “star-ground” as an approximation of this network. It is imperfect for two practical reasons: First, the star-ground “terminal” is not a “point” but a short loop (albeit heavy copper wire, as noted in Section 4.2). Second, in contrast to what my schematics might imply in an effort to avoid clutter, some “chain” grounding is used within channels (or tube stages) over short distances in cases where relatively low current is expected. These “sub-stars” originate at the negative terminals of decoupling or filter capacitors serving that stage or channel. You may visualize this as a hierarchical star grounding rule in the modified vintage unit.

4.5. Power Supply: Auxiliary Panel. As shown in Figure 6, the auxiliary panel's power supply has three sets of bi-polar outputs: ±18 V for the balanced line drivers in CH1-CH4, ±15 V for the associated clip alert circuits, and ±12 V to control relays in the balanced input signal path of CH1-CH4. This power supply was built on a perf-board using point-to-point wiring, with attention to heavy and redundant ground conductors leading to the star ground terminal at the bottom of the board. A photo of the installed board is shown in Figure 7.

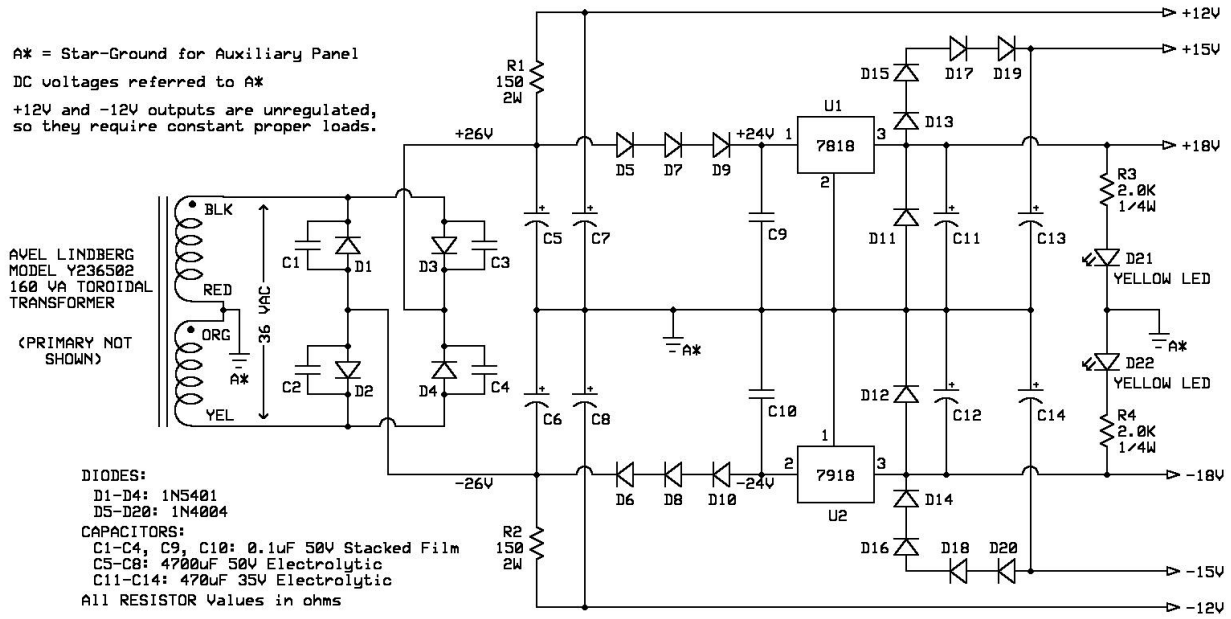


Figure 6. Schematic diagram of power supply in the modified Altec 1567A's auxiliary panel. See Figure 4 for transformer's primary circuit.

The secondary windings of the toroidal power transformer, each rated nominally 18 V_{RMS}, are hooked in series for 36 V_{RMS} output with a grounded center tap. D1-D4 form a full-wave bridge rectifier with bi-polar DC outputs referenced to ground and filtered by C5 and C6. Bypassing each rectifier diode, C1-C4 are snubber capacitors intended to silence RF switching noise.

The ±12 V outputs are unregulated and depend on the voltage drops across R1 and R2; they are filtered to low ripple by C7 and C8. Correct voltage here requires invariable loads in the relay control circuits served. As detailed in Section 4.6, six of the relays (those of CH1 and CH2) operate on +12 V while the other six (for CH3 and CH4) use -12 V; each relay control switch routes current to either a relay or an equivalent dummy load resistor to maintain constant current in the ±12 V circuits.

The regulated outputs (±18 V and ±15 V) use heat-sink-mounted one-amp voltage regulator ICs U1 and U2 (LM7818 and LM7918). Each of them has three diodes (D5-D10) in series with their input, which drops input potential by about 2.1 volts and diverts some unnecessary heat dissipation away from the regulators. Physically close to the input pins, C9 and C10 are bypass capacitors recommended when using three-terminal voltage regulators. At their outputs, D11 and D12 help protect the regulators in the unlikely event of back-EMFs produced by the loads,

and C11 and C12 help filter out noise in the regulator outputs. Of course, the power supply's ±18-V outputs come directly from the regulators; yellow LEDs D21 and D22, mounted on the auxiliary panel to the left of the main power switch, indicate power-on status. To get the ±15-V outputs, each regulator's output potential is dropped through a series of four diodes (D13-D20) then filtered by C13 and C14.

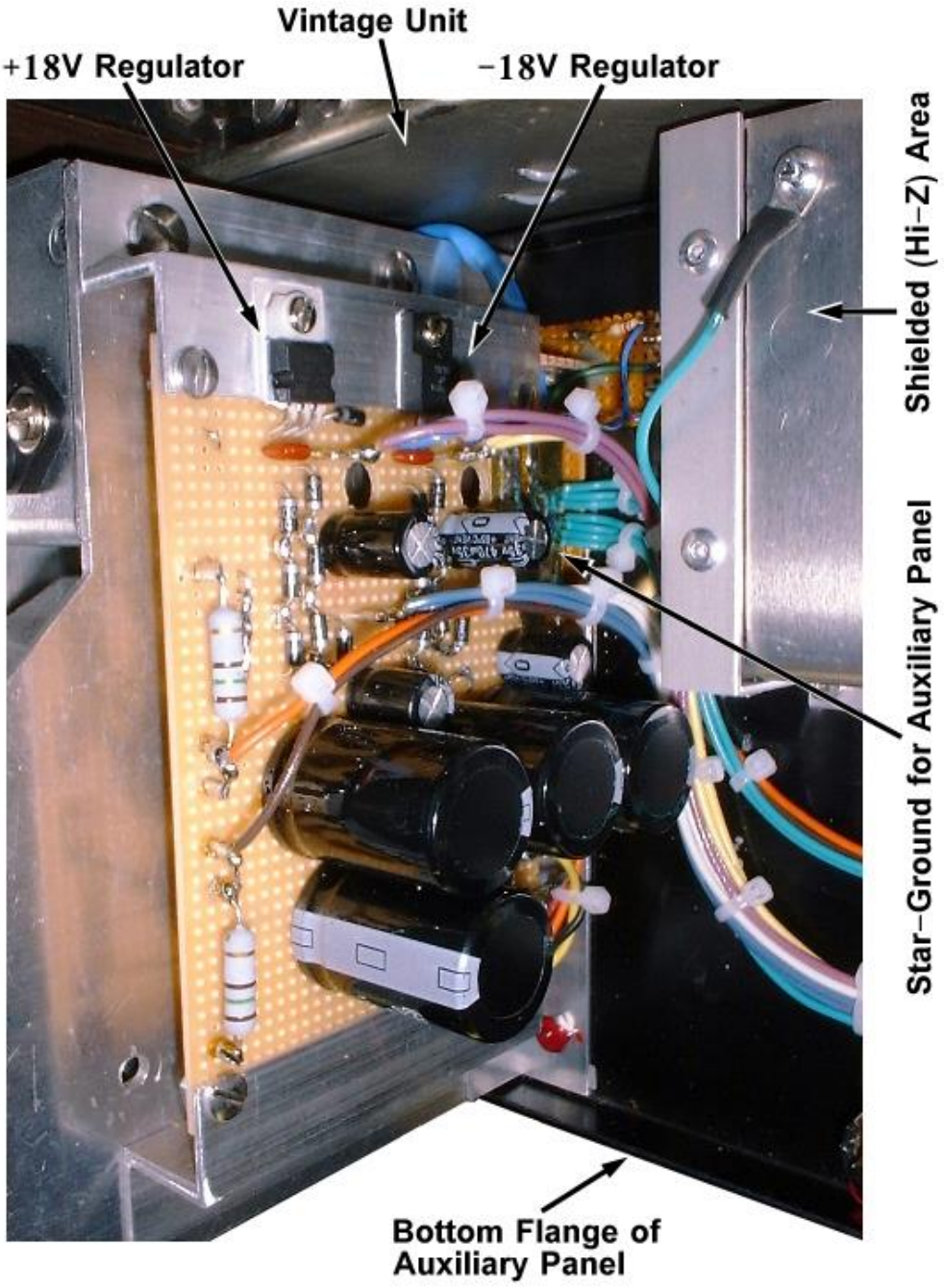


Figure 7. Labeled photo of installed power supply board for auxiliary panel.

4.6. Channels 1-4: Balanced Input Circuits. Linking the female XLR jacks to the input transformer primaries, CH1-CH4's balanced input circuits are almost identical, so the schematic in Figure 8 includes just one channel's signal path to minimize clutter. For each channel, three Omron G5A-234P DPDT relays (RY1-RY3) execute pad, polarity and impedance options; along with associated components, six relays occupy each of two perf boards mounted adjacent to input transformer sockets in the vintage unit. Figure 9 is a view inside the vintage chassis showing these boards.

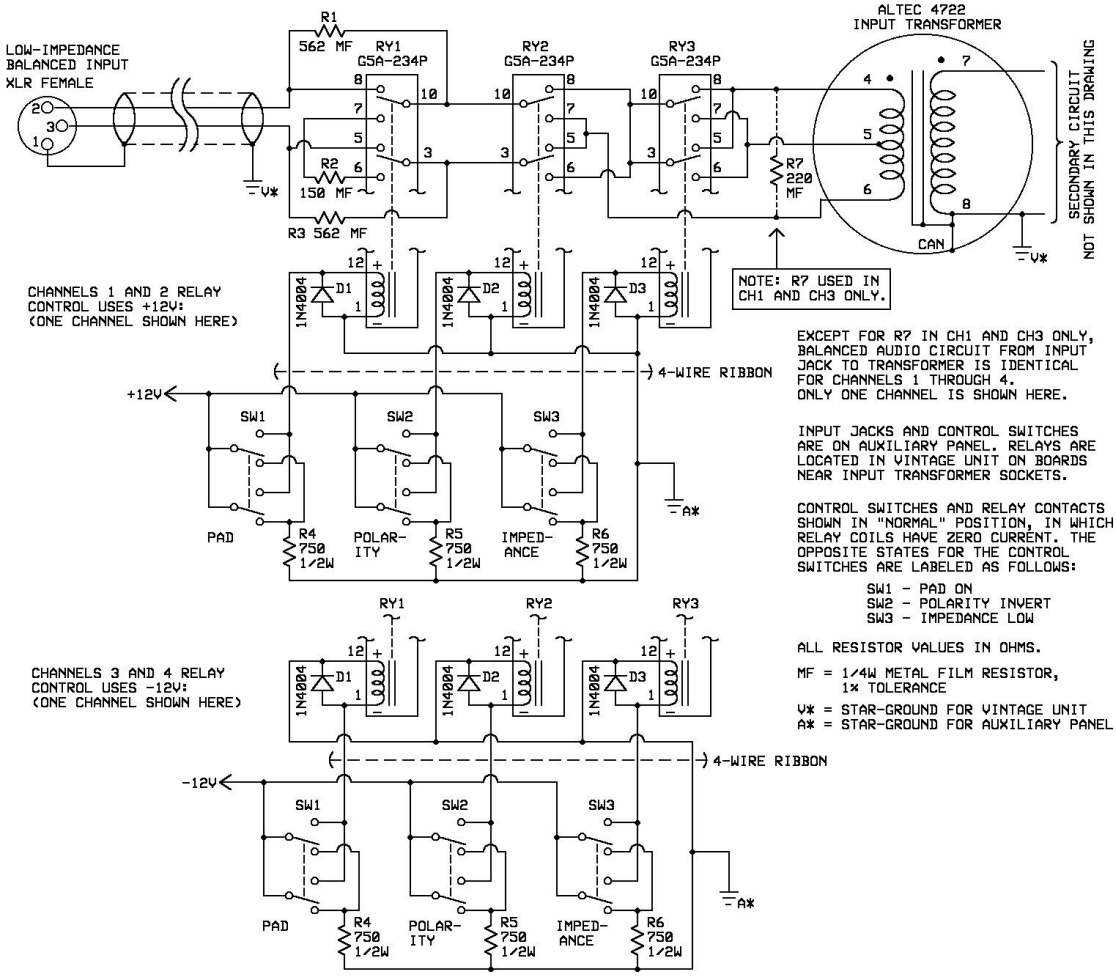


Figure 8. Schematic diagram of balanced input circuits for CH1-CH4 of the modified Altec 1567A. Except for R7 as noted, the balanced signal paths for these channels are identical, so only one is diagrammed here (top). Depending on channel, relay control circuits use either 12 VDC polarity, and each is shown (middle and bottom).

Working left-to-right across the top of Figure 8, I'll focus first on the signal path before describing the relay control circuits. Female XLR jacks located on the auxiliary panel are wired to the relay boards using shielded twisted-pair cable. In its "normal" (off) state, relay RY1 simply passes the balanced signal, but its activation engages balanced "U-pad" resistor network R1-R3. Nominal pad loss is -20 dB; however, this (and how input Z changes) depends on the channel, as

explained in Sections 5.2 and 5.3. RY2 inverts the balanced signal when activated; normally the balanced output jacks for CH1-CH4 have the same polarity as their input jacks. With its contacts wired in parallel as SPDT, RY3 normally connects the balanced signal to the input transformer's full primary coil (pins 4 and 6); RY3 activation switches to the tap (pin 5) to reduce input Z.

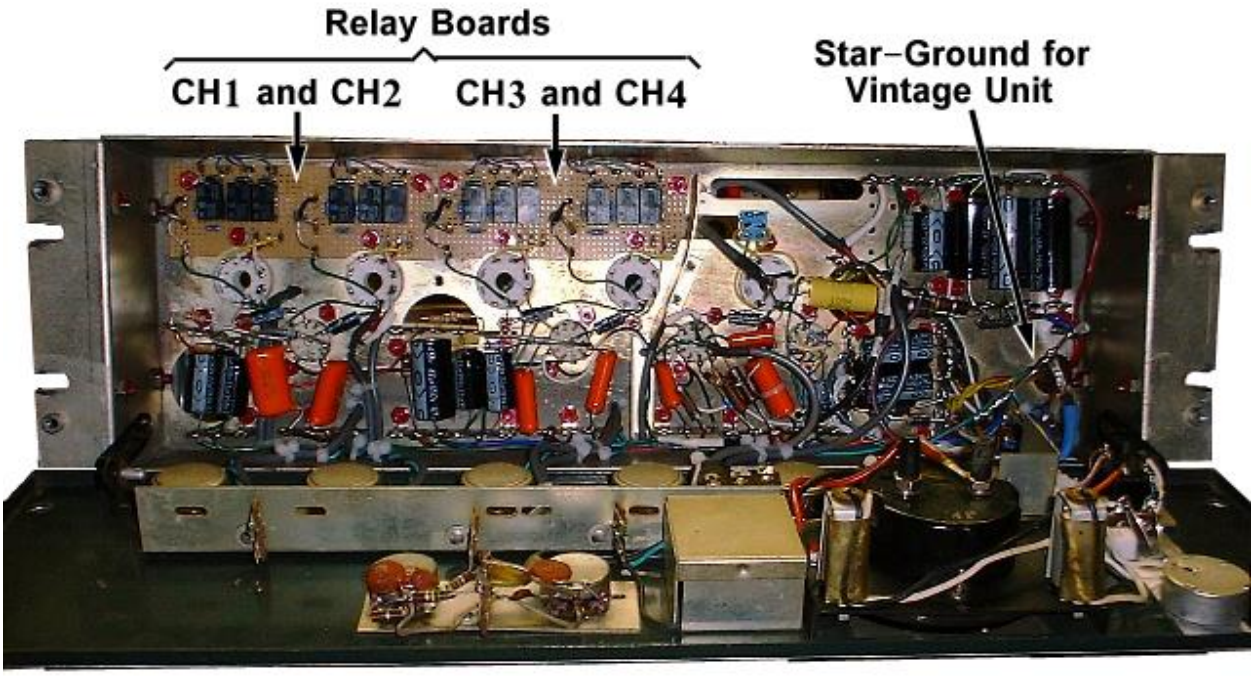


Figure 9. Photograph of interior of vintage unit (its front panel swung open) of the modified Altec 1567A. Relay boards (upper left) place relays near transformer sockets.

Up to this point in the description, the balanced input circuits of CH1-CH4 are identical. Now comes the place where they differ: the input transformers' full primary winding is shunted by 220-Ω resistor R7 in CH1 and CH3, but not in CH2 and CH4. Note that channels containing this shunt are comparable to the original Altec1567A design (see Appendix), which places a 180-Ω primary shunt resistor across each input transformer. Availability of un-shunted inputs increases channel diversity, as explained further in Section 5.2.

Each relay is linked to a control switch mounted on the auxiliary panel. For each of CH1-CH4, these are SW1-SW3 (DPDT mini-toggles wired SPDT), respectively controlling RY1-RY3. "Normal" switch handle positions are "down," corresponding to inactive relay coils. The relays operate at 12 VDC; to balance the load on the auxiliary panel's bi-polar power supply, relay control for CH1 and CH2 uses +12 V (middle region of Figure 8), while that of CH3 and CH4 uses -12 V (bottom region of Figure 8). "Normal" settings of SW1-SW3 engage 750-Ω dummy load resistors R4-R6, respectively. This resistance is equivalent to a relay coil, so load on the ±12 V supplies remains nearly constant regardless of the combination of relays activated. Shunting relay coils, diodes D1-D3 suppress back-EMF spikes when relays change state.

4.7. Channels 1-4: Triode Stage. The triode circuits in CH3 and CH4 are essentially identical to those of a stock Altec 1567A (see Appendix), with the omission of the wire from the pre-fader

output to pin 1 of the transformer socket (this was a negative feedback connection needed only with the “phono equalizer” plug-in accessory). However, the grid circuits of CH1 and CH2 are built out in shielded compartments of the auxiliary panel so that each includes a high-Z unbalanced input, pad, and variable attenuator. Each alternative grid circuit is included in Figure 10; I will describe the common aspects of the triode gain stage first, followed by detailing the grid-circuit elaboration specific to CH1 and CH2.

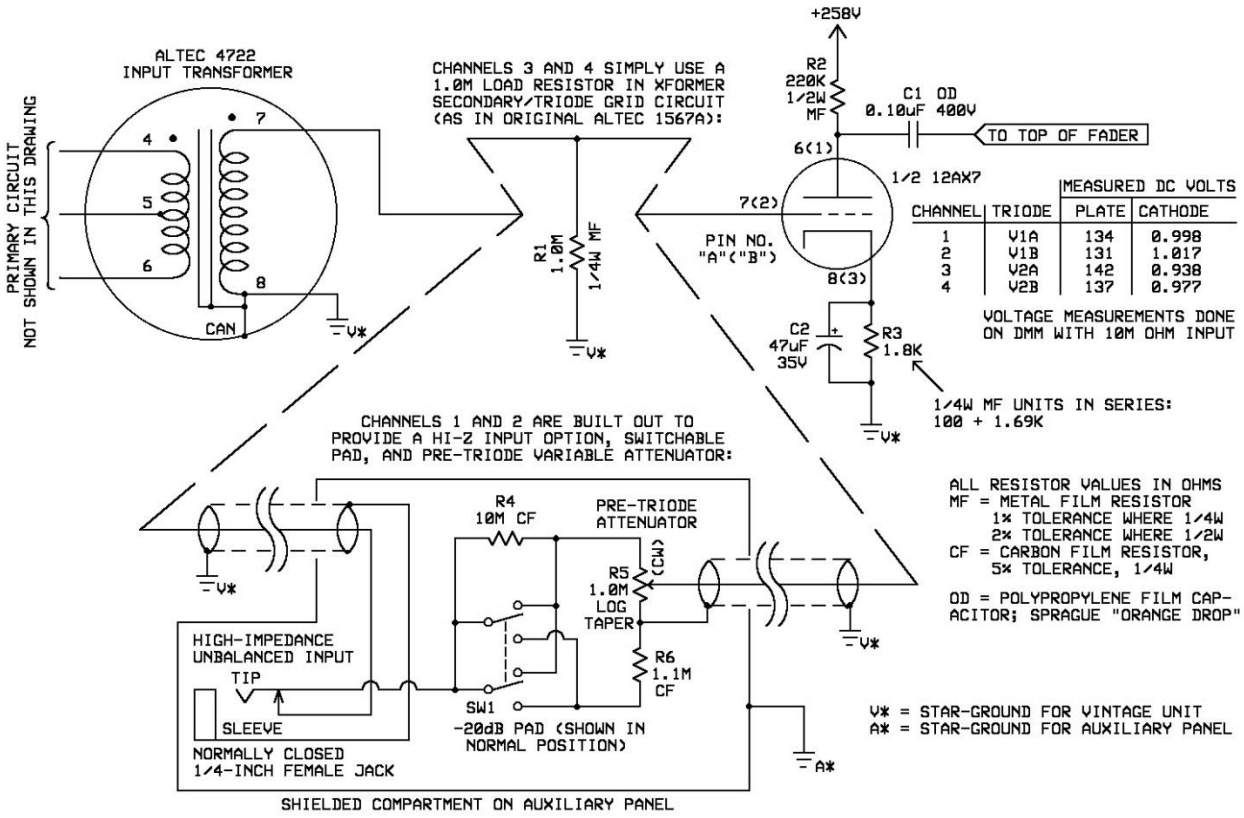


Figure 10. Schematic diagram of triode stage used in CH1-CH4 of modified Altec 1567A. CH1 and CH2 have a high-impedance (grid) circuit build-out that includes an unbalanced input jack, attenuator pot, and pad switch (bottom of diagram). Instead, CH3 and CH4 simply use 1-MΩ resistor R1, retaining the vintage design. The dashed lines depict the placement of the two alternative high-Z circuits.

The nominal 50-KΩ secondary winding of the Altec type 4722 input transformer needs a load resistor to reflect the proper impedance to the primary circuit; this is provided by 1-MΩ resistor R1 (Figure 10) for CH3 and CH4 (exactly like the stock unit), or the high-Z build-out for CH1 and CH2 (also representing a 1-MΩ resistive load; see below). At 1 KHz, voltage step-up in the input transformer is about 25 dB when the full primary winding is used. The signal at the secondary is applied to the grid of one triode of a 12AX7 (either V1 or V2). Plate load resistor R2 and cathode bias resistor R3 have the same values as their counterparts in the original design. This gives a triode operating point close to the original design’s (measured DC potentials on the plate and cathode listed in Figure 10 are within about 10 percent of those given in the original schematic).

As in the original, C2 shunts the cathode bias resistor to eliminate negative (degenerative) feedback at audio frequencies, maximizing gain at the cost of some bandwidth loss.

As in the original design, C1 AC-couples the output of the triode stage to the “top” (clockwise-most) terminal of the channel’s output fader pot (see Figure 12, R1). High-quality SBE (formerly Sprague) 716P-series “Orange Drop” polypropylene film capacitors are used for signal coupling throughout the modified vintage unit (with one exception noted in the CH5 description, Section 4.10). Voltage gain in this triode stage (loaded by a 250-KΩ fader pot) is 35 dB. Accounting for transformer step-up (25 dB using the full primary coil as noted above), transformer plus triode gain is therefore about 60 dB (see Section 5.5 for details of CH1-CH4 gain structure). Note that common-cathode gain stages such as this invert the input signal.

Instead of R1 to load the high-Z link between the transformer and triode, CH1 and CH2 are each provided a normally-closed unbalanced input jack, pad, and variable attenuator, as diagrammed in the bottom portion of Figure 10. Shielded cables to and from this network were made as short as possible. Figure 11 shows the inside of the auxiliary panel’s shielded area, its cover removed, with close-up images of the high-Z build-outs (see Section 4.10 for a description of CH5’s). The shield cover is designed to be detached and removed (with care) without disturbing the cables running into or behind the shielded area, in case service or modification of the enclosed circuitry is needed.

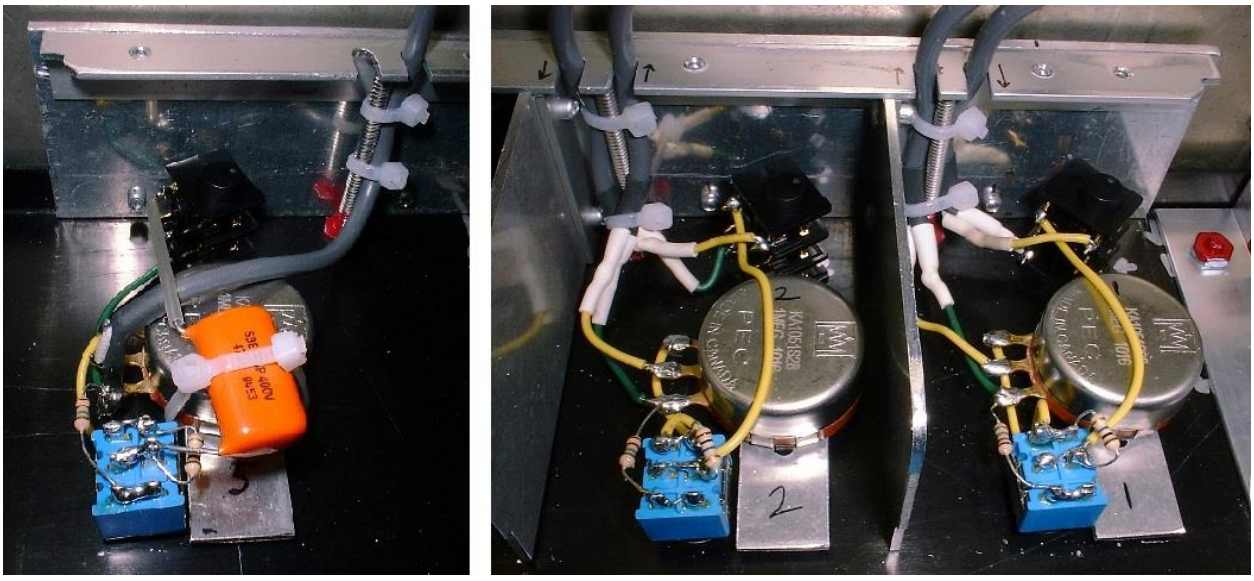


Figure 11. Photos of the high-impedance circuitry inside the shielded compartment of the auxiliary panel. The left image shows CH5’s build-out, while those of CH2 and CH1 are in the right image. The two interior shields separating the channels are visible in the right image. Each pot’s mounting bushing and anti-rotation tab pass through a rectangular aluminum piece, the bottom of which engages the shield cover when installed.

A Neutrik unit with gold-plated contacts, the ¼-inch female jack for the high-Z unbalanced input of CH1 and CH2 has normally closed contacts linking the input transformer’s signal to the pad/attenuator network. Inserting a plug opens that connection and substitutes the plug’s signal.

SW1 is a DPDT mini-toggle switch wired as SPDT for enhanced reliability; its normal setting (handle “down”) applies the signal directly to the top (CW or clockwise-most) terminal of “pre-triode attenuator” pot R5, whose wiper is hooked to the triode’s grid. This is a robust mil-spec 1-MΩ log-taper pot from Precision Electronic Components, Ltd.; its operating voltage limit is 500 V_{RMS}, which should be sufficient to withstand distortion experiments using transformer saturation (see Sections 5.11, 5.12 and 5.13). In SW1’s handle-“up” setting (pad engaged), R4 is placed in series with pot R5, and R6 shunts the R4 + R5 series. The result is a -20 dB loss while keeping the resistive load presented by the network near 1 MΩ. However, capacitive reactance in the triode lowers the network’s input impedance to about 840 KΩ at 1 KHz for the normal settings (attenuator full clockwise and pad off; see Section 5.4 for more details).

If you are concerned about the high-Z build-out network of CH1 and CH2 causing high frequency loss due to stray capacitance, your concern is justified. Fortunately, as I will show in Section 5.13, the most useful settings of the pre-triode attenuator pot (fully clockwise, or the region near fully counter-clockwise) are not adversely affected. But users of this experimental gear need to be aware of this.

4.8. Channels 1-4: Balanced Line Drivers. The solid-state differential output driver and clip alert indicator circuits used by CH1-CH4 are diagrammed in Figure 12. Only one of the four identical driver/indicator channels is shown; the driver stages (using ICs U1 and U2) are in the top half, and the indicator circuit (using U3 and U4), the bottom half of this schematic. This circuitry is built on two perf-boards (one for CH1/ CH2, one for CH3/CH4) which are mounted to an aluminum support plate adjacent to the output jacks on the auxiliary panel. The support plate doubles as a shield to help isolate these outputs from the adjacent low-Z input jacks. A photo of the installed boards, and a close-up of the CH1/CH2 board, is shown in Figure 13. After noting some general aspects of these boards, I will discuss the driver circuit later in the present section, and then the clip-alert indicator in Section 4.9.

The output driver circuit uses ±18 V power supply rails, and the clip alert circuit ±15 V rails. Electrolytic 10-µF capacitors C1 and C2 bypass the ±18V rails at power supply connections to each board, as do C10/C11 at the ±15V connections. The DC supply pins of each IC package are bypassed with 0.1-µF capacitors (e.g., C5 and C6 for U2). For minimum inductance at these bypasses, stacked-film capacitors are used with absolute minimum lead lengths; the capacitor and IC pins share the same perf-board hole at each IC’s power connections.

For quietest and most stable performance, the balanced output driver circuits occupy areas of the perf-boards that were prepared with solid ground planes. Since this meticulous construction technique may be unfamiliar, I will summarize it: First, the component layout is carefully planned for minimum point-to-point interconnection distance beneath the board. Second, using 0.1-inch-grid graph paper, an actual-size template for the ground plane is designed to cover a maximum continuous area; it contains holes to give non-grounded component pins free access to their intended perf-board holes. Third, 0.005-inch-thick copper foil is cut to match the template, sanded lightly on the bottom surface, and applied to the top of the perf-board using heat-tolerant epoxy cement. Finally, during assembly, grounded component pins are splayed out and then soldered directly to the ground plane; the non-grounded pins, which pass through the plane, are

soldered point-to-point beneath the board. Careful attention to layout and construction yields performance rivaling a well-designed double-sided PCB.

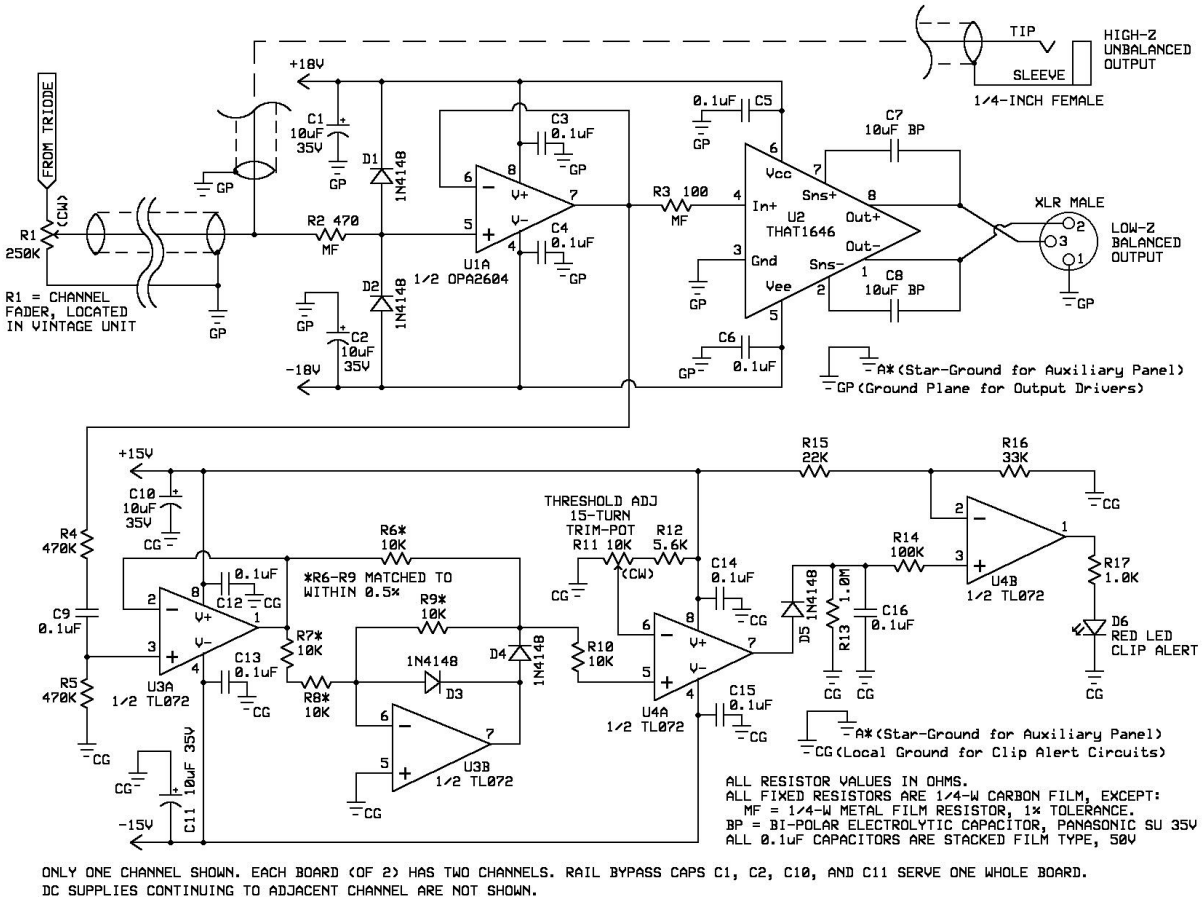


Figure 12. Schematic diagram of CH1-CH4's solid-state line driver and clip alert indicator circuits; channel fader and unbalanced output are also shown. These circuits are identical for CH1-CH4, so only one is diagrammed here.

At the upper-left of Figure 12 is pot R1, a channel fader in the vintage unit (for either CH1, CH2, CH3, or CH4); the signal is taken directly from its wiper via a shielded cable leading to an output driver board. Importantly, the fader pots are grounded to the driver board's ground plane, not to the vintage unit's ground, to preclude an internal ground loop. At the connection to the driver board, a second shielded cable branches the signal to a ¼-inch female high-Z unbalanced output jack; the other branch feeds the balanced output driver.

The THAT1646 differential output driver chip (U2) has a relatively low input impedance of 5 KΩ, so it needs to be fed by an op-amp stage. This is the role of U1A, which is one-half of a Burr-Brown (now part of Texas Instruments) OPA2604 dual op-amp (the other half is used in the board's other channel). The OPA2604 is an excellent audio performer, with very low noise and distortion for a high-impedance FET-input device; here it is configured as a non-inverting unity-gain buffer (voltage follower). In series with its input, R2 is meant to assure stability of this stage; fast-acting diode clamps D1 and D2 protect U1A from damage when peak signal

voltages exceed +18 V or drop below -18V. Note that under such conditions, clamping is not isolated from the high-Z unbalanced output and distortion will show up there (as well as at the balanced output). This matter is discussed in Sections 5.8 and 5.14.

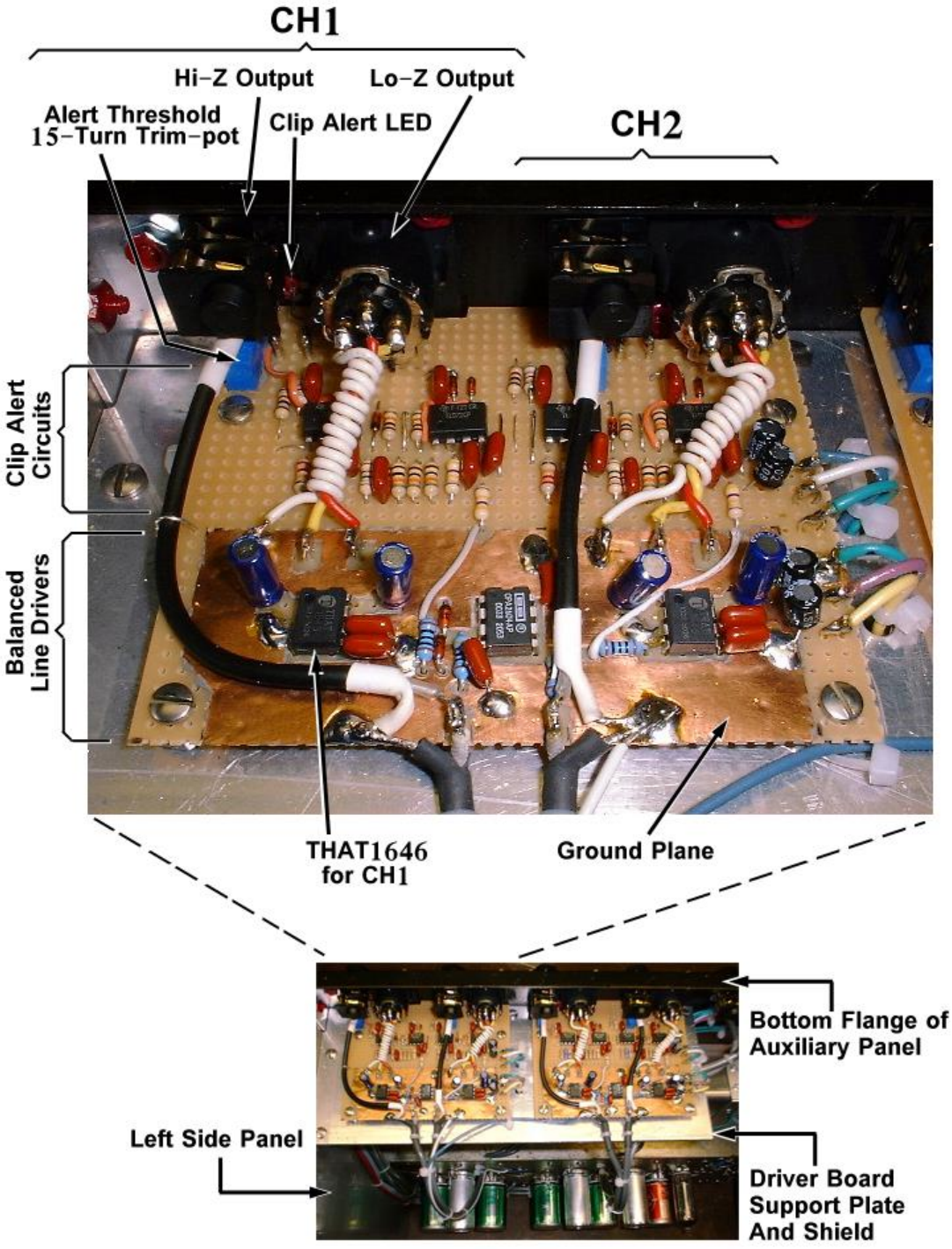


Figure 13. Labeled photos of CH1-CH4's solid-state output driver boards. Looking up toward the underside of the modified unit, the bottom photo shows how both boards mount to the assembly. The top photo is a close-up of the CH1/CH2 board.

U1A's output branches to the clip alert circuit (described below) via R4 and to the balanced output driver IC (U2) via R3. The THAT1646 is implemented exactly as set forth in Figure 5 of That Corporation's spec sheet and applications guide for this IC (their Document 600078 Rev 04; PDF available on-line). Non-polarized ("bi-polar") electrolytic capacitors C7 and C8 AC-couple feedback in the driver's servo loops to minimize DC offset at its outputs. The THAT1646 emulates transformer-balanced outputs in many respects; I will describe its characteristics more completely in Section 5.9.

Note that U2's "+" (non-inverted) output feeds pin 3 of the male XLR output jack, and the "-" (inverted) output feeds pin 2. You may initially think this conflicts with the conventional standard of using XLR pin 2 for "+" or "hot"; however, remember that the channel's triode stage inverts the signal, so this crisscrossed output connection makes the balanced output jacks for CH1-CH4 match the polarity of their respective balanced inputs (and the high-Z inputs of CH1 and CH2). However, the high-Z unbalanced output remains inverted with respect to standard XLR polarity (and the high-Z inputs of CH1 and CH2). Of course, setting a polarity switch to "invert" affects only balanced input signals, not the high-Z inputs.

4.9. Channels 1-4: Clip-Alert Indicator Circuit. The clip alert indicator (Figure 12, bottom half) works by detecting whether a pre-set instantaneous (peak) amplitude is exceeded at the balanced line driver IC's input. Powered at ±18 V, the output of U1A and individual output pins of U2 (i.e., measured in single-ended mode, not differential mode; double this for differential mode) can swing between about +15 V and -15 V without clipping. Importantly, when feeding U2, U1A has slightly more headroom than U2 itself; also, U2's clip threshold decreases as output loading increases, as described in Section 5.9. Thus, by monitoring the output of U1A, the clip alert circuit is assured of receiving a clean signal when the balanced channel output has just exceeded its clip threshold; threshold adjustment range of the clip alert therefore brackets all conditions in which clipping at the main outputs can occur.

The clip alert uses inexpensive yet effective TL072 dual FET-input op-amps U3 and U4, run at ±15 V. Setting the circuit's input Z at nearly 1 MΩ, voltage divider R4-R5 halves the signal's amplitude (i.e., makes a -6 dB pad) so it can never exceed U3A's headroom. C9 blocks any DC offset of U1A's output (which is no greater than a few millivolts). Working as a unity-gain buffer, U3A drives an absolute value circuit built around U3B. This is literally a "cookbook" circuit by Walter G. Jung ("IC Op-Amp Cookbook" 3rd Ed., 1986, ISBN: 0-672-22453-4; p.245, Fig. 5-14B), to whom I refer you for a detailed description. Basically, like a full-wave rectifier, it inverts only the signal's negative voltage swings and doesn't affect the positive ones. This lets the following stage (threshold detector) respond to both positive and negative peaks of the original signal.

Configured as a voltage comparator, U4A serves as the threshold detector. It compares the absolute value signal at its non-inverting input to a reference voltage at its inverting input. The reference voltage comes from the wiper of 15-turn trim-pot R11, set up as a variable voltage divider; series resistor R12 scales R11 to a maximum useful adjustment range. Accessible through the front of the auxiliary panel next to the associated clip alert LED, this trim-pot sets the amplitude at which the LED activates. Clockwise rotation increases the threshold.

Comparators lack feedback, so full open-loop gain holds U4A's output at its maximum negative level until a super-threshold peak is detected, and then the output swings positive. In series with U4A's output, D5 passes only positive voltages on to LED driver stage U4B, also wired as a comparator, but with a fixed reference voltage determined by R15 and R16. The input network of U4B includes C16, which charges rapidly through U4A's low output impedance when a peak is detected. Afterwards, C16 discharges relatively slowly, mostly through high-value resistor R13 since U4B's input impedance is very high. This extends the duration of transient peaks enough to make their resulting LED flashes visible.

Setting CH1-CH4's clip alert thresholds requires monitoring the outputs with an oscilloscope; I will describe a calibration procedure in Section 5.10. Before shipping the modified Altec 1567A, I adjusted all channels to indicate clipping of balanced outputs into 600-Ω floating loads.

4.10. Channel 5. For CH5, the old Altec 1567A master channel was isolated by substituting an unbalanced high-Z input network for the original mix bus, among other modifications. The modified channel is diagrammed in Figure 14. Comparison to the original schematic (see Appendix) may be helpful as the following description emphasizes the modifications.

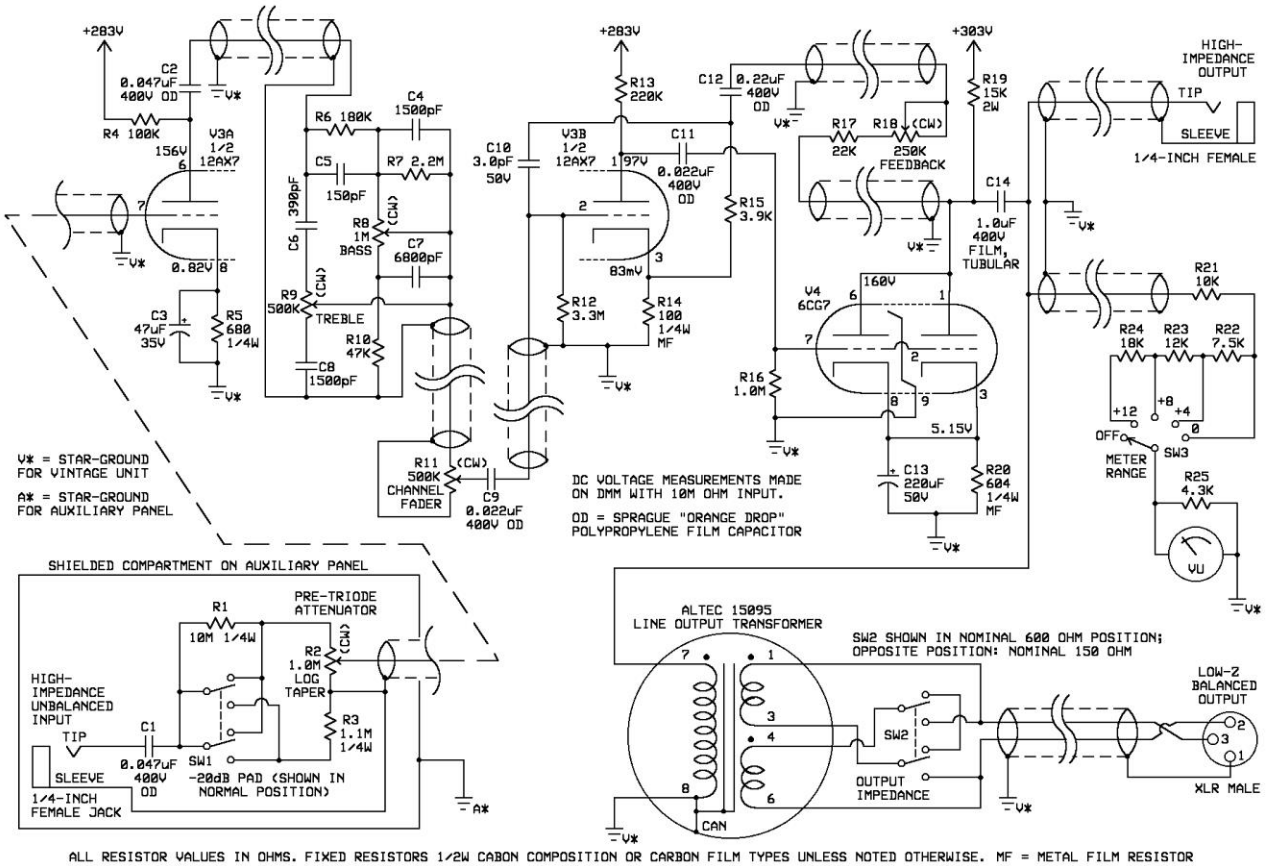


Figure 14. Schematic diagram of CH5 in modified Altec 1567A.

As shown in the lower left corner of Figure 14, the ¼-inch unbalanced input jack occupies one compartment of the auxiliary panel's shielded high-Z area (see left-hand photo in Figure 11).

This input is AC-coupled by C1 to a switchable pad/variable attenuator network like the ones used in CH1 and CH2 (see Figure 10 and its description in Section 4.7). As in those channels, this network interacts with the following triode to set CH5's input Z to about 840 KΩ at 1 KHz for the normal attenuator and pad settings (full clockwise and off, respectively). Also as in CH1 and CH2, stray capacitance restricts the most useful range of pre-triode attenuator R2 to its relatively extreme settings (and only the counter-clockwise region with the -20-dB pad on). Normal settings for most applications should be pad off and attenuator fully clockwise; see Section 5.13 for more information.

Via a minimum length of shielded cable passing into the vintage unit, the signal at the wiper of R2 is applied directly to the grid of V3A, one triode of a 12AX7. Unlike the corresponding triode's hook-up in the original unit (the mix bus amplifier or summing mixer), here V3A uses cathode bias provided by R5. Compared to the original design, this should increase the maximum input amplitude this stage can handle. Electrolytic capacitor C3 bypasses R5 to prevent degenerative feedback and maximize gain. The chosen value of R5 (680 Ω) optimizes the triode operating point while retaining plate load resistor R4's original value (100 KΩ); this should keep this stage's output Z similar to the original's for driving the tone control/channel fader network which follows.

The original pre-tone-network "recorder output" connection (see Appendix) is omitted in the modification. The tone control network itself, including original bass and treble pots R8 and R9, is one of the two vintage sections that were left original and not re-built completely during the modification (the other is the VU meter network). It thus retains the original ceramic disc capacitors C4-C8 (such capacitors age better than film capacitors used for DC-blocked signal coupling) and carbon composition resistors R6, R7, and R10 (these read within tolerance on an ohmmeter). The tone network's output feeds channel fader pot R11 via the original short length of shielded cable. At the fader's wiper, coupling capacitor C9 is a new SBE "Orange Drop" 716P-series polypropylene film-and-foil unit (as are all coupling capacitors throughout the modified vintage unit, except C14 in this channel; see below).

From C9, the post-fader signal passes via shielded cable to the grid of triode V3B, which is the first stage of CH5's two-stage output driver (called a "line amplifier" in the original Altec 1567A manual; see link in Appendix). This stage is re-built exactly like the original, and no attempt was made to "correct" an observed minus-twenty-percent difference in measured plate voltage here compared to that published in the original schematic (97 V here versus original 120 V; see Appendix). This was the largest plate voltage difference versus the original schematic.

The final vacuum tube stage uses the two triodes of V4, the 6CG7, hooked in parallel to act like a single triode with lower output impedance. (Note that the original schematic omits the 6CG7's internal shield at pin 9; although unimportant in this application, it was grounded in the original unit, as it is in the re-build as shown in Figure 14.) The original cathode bias resistor (R20) value of 470 Ω caused V4 to conduct a little too much current in my opinion (10.7 mA versus the 9.33 mA deduced from voltages in the original schematic); this made plate load resistor R19 dissipate about 1.7 W, closer to its 2W rating than was comfortable (R19 is one of the cases where I recycled the original component; like me, I wanted it to spend its golden years doing a

little less work than perhaps it had in the past). Increasing R20 to 604 Ω made V4's plate and cathode voltages more closely match those given in the original schematic, and made R19 dissipate a cooler 1.36 W. It's unclear whether this bias tweak was prompted by an aging 6CG7 or an "outlier" (I had no other 6CG7 for comparison), or some other cause; but when it's time to replace V4, please re-check the new one's voltages and re-bias if necessary.

The stock vintage unit used a 47 KΩ fixed resistor (R32 in Appendix) to set feedback between the output driver's two triode stages (V4 and V3B). Instead, I used an interpretation of an adjustable-feedback modification on an Altec 1567A mentioned by Eddie Ciletti in his December 2010 Mix magazine "Tech's Files" column (pages 60-62). My version wires the old "mix 5" 250-KΩ pot (the original passive input channel fader, R18 in Figure 14) as a rheostat to replace the fixed 47-KΩ feedback resistor. At 22-KΩ, R17 limits the maximum negative feedback allowed (when R16 is full counter-clockwise). On the vintage panel, I marked the feedback knob position equivalent to the original fixed resistor (the "design feedback level"); more clockwise settings decrease feedback, thus increasing gain and harmonic distortion, compared to the original design, and vice-versa for more counter-clockwise settings (see Section 5.17). Aside from the shielded cables to the front panel necessary to patch in R17 and R18, other components completing the feedback loop are like the original design; C12 blocks DC, C10 may be for stability and/or to tweak high-frequency response, and R15 couples feedback to V3B's cathode.

As in the original unit's master channel, V4's output is AC-coupled using a 1.0-µF capacitor (C14 in Figure 14). This is the only non-SBE new coupling capacitor in the modified vintage unit; it's a generic metalized-film tubular unit sold by Antique Electronics Supply. From there, the signal branches to the output transformer primary and, via shielded cables, to the ¼-inch unbalanced ("high Z") output jack on the auxiliary panel and the VU meter network on the vintage panel. (The original meter circuit was not re-built, however its ground lead was segregated from that of the tone network, which it had originally shared.)

Accessible on the back of the vintage chassis next to the output transformer socket, DPDT mini-toggle switch SW2 was added to easily set CH5's nominal balanced output Z for either 600 Ω (secondary windings hooked in series) or 150 Ω (parallel hookup). Via a shielded twisted-pair cable, the auxiliary-panel-mounted XLR male balanced output jack is wired such that its pin 2 has the same polarity as the channel's unbalanced input. But note the unbalanced output has the opposite polarity since this channel uses an odd number (three) of common-cathode (thus inverting) triode stages.

5. Performance and Applications of Modified Unit

5.1. Standard Amplitude Units. Most of this report expresses signal amplitudes in dBV, or decibels referred to 1 V_{RMS} (i.e., 0 dBV = 1 V_{RMS}) for ease of calculation. When gear to be compared or interfaced specifies amplitudes in dBu or dBm, conversion may be desired. Since dBu is referred to 0.7746 V_{RMS}, 0 dBV = 2.22 dBu; just subtract 2.22 from any dBu value to get dBV, or add 2.22 to a dBV value for dBu. Units of dBm were most frequently used for vintage gear, where 0 dBm = 1 mW (milliwatt); conversion to dBV or dBu requires knowing the load

impedance or resistance. Conversion is trivial for 600-Ω loads because dBm = dBu in that particular case, and it is the most common case. The original Altec 1567A manual (see Appendix for link) uses dBm; Section 5.21 of this report expresses CH5's output in dBm for easy cross-reference to that manual. "Volume Units" (VU), which include aspects of signal dynamics, are not necessarily converted easily into decibel units (see Section 5.21).

5.2. Channels 1-4: Impedance of Balanced Inputs. In the classical era of audio engineering when the Altec 1567A was designed, optimum signal transmission from a source (such as a mic) to an input (as in a preamp channel) was usually assumed to need maximum power coupling. This calls for equal source and input impedances, where voltage at the input terminals equals that dropped across the source's internal impedance—a 6-dB voltage loss. In our modern era, audio gain models use voltage terms (at least in preamp stages), so impedance-matched transmission lines seem inefficient to us. For example, today's mic inputs commonly have impedances some 10-fold higher than the mics they host (this is called a "bridging" connection), for a much smaller voltage loss across the mic's internal impedance. However, I don't necessarily favor throwing out yesterday's audio traditions for the sake of a few dB of efficiency; such decisions can have trade-offs. Source-preamp interaction involves a complex interplay of reactive elements when transformer-coupled tube stages are used. This adds up to tonal "character," which hopefully is often good. Some of the sought-after "vintage tone" may well depend on maximum power coupling and impedance matching.

By taking advantage of an unusual feature of the stock Altec 1567A's microphone input circuits (and one irregular input transformer), I opted to give the modified version some diversity of balanced input impedances among channels 1-4. Users can then experiment and decide for themselves what sounds best in different cases. Notice in the original schematic (Appendix) that each channel's input transformer primary was shunted with a 180-Ω resistor. An uncommon strategy for tube preamps, this shunt accounts for most of the nominal 150-Ω impedance of the mic inputs (with the "normal" input-Z connection, which uses the full primary winding). I omitted this resistor in CH2 and CH4 but included a 220-Ω resister in CH1 and CH3 (R7 in Figure 8). (Why use 220 Ω rather than the very similar 180 Ω? A lack of will to go the distance, perhaps.)

As mentioned in Section 2.2, the model 4722 transformer originally in CH3 tested differently than the others. While its DC resistance readings, frequency response, and voltage step-up ratio matches the others, its input impedance is inherently low. In impedance tests, it acts like it has an internal shunt resistor. At any rate, I moved this transformer to CH4 in the modified preamp. Users should be alert to CH4's irregular (and possibly defective) input transformer while taking advantage of its different impedance.

To evaluate the impedance of each balanced input, I used the millivolt meter in a Hewlett-Packard 331 Distortion Analyzer (which has a 1-MΩ input) to measure RMS voltage drops across a 1.0-KΩ resistor in series with the channel input. Low-amplitude test signals were sine waveforms of various frequencies from a function generator (B&K Precision 3011B), whose verified 50-Ω output Z was factored into the calculations. Note that this technique gives only the magnitude of the impedance vector, not its angle (i.e., the proportion of resistive and reactive components is not determined).

In the case of impedance and pad switches set for “normal” (handle down), the results for CH1-CH4 are plotted in Figure 15. For each channel, input impedance peaks in the audio mid-band; this is most dramatic in CH2, which lacks the primary shunt resistor, and impedance reaches nearly 2 KΩ at 1 KHz. CH4’s “irregular” transformer also lacks the shunt, but this channel has a broader impedance peak which reaches only 390 Ω. By virtue of their shunt resistors, CH1 and CH3 probably most closely reconstitute stock Altec 1567A mic input channels; in the modified unit they display a fairly broad impedance peak reaching 185 Ω. Sagging input impedance with decreasing bass frequency presumably comes from the non-ideal characteristics of these (real) transformers, such as reactance from the transformer’s self-inductance. Impedance drop-off with increasing treble frequency is probably largely due to the triode stage’s capacitive reactance (Miller Effect; see Section 5.4) reflected to the transformer’s primary.

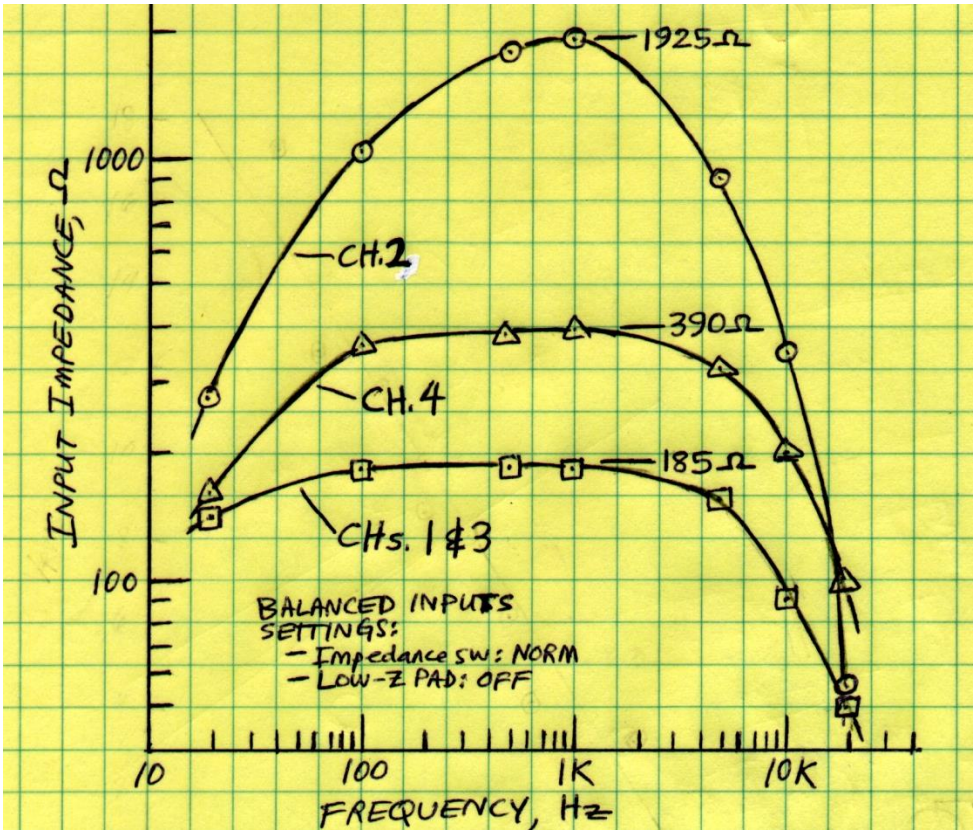
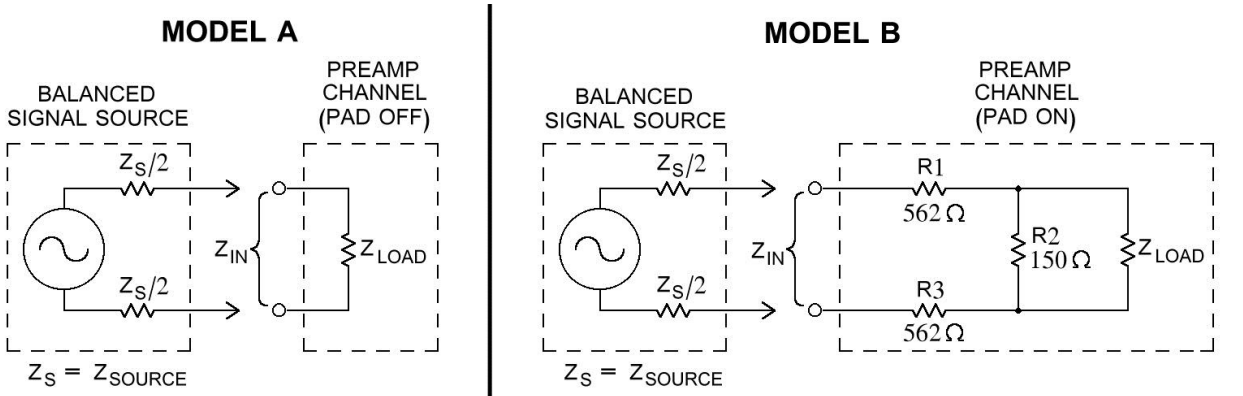


Figure 15. Measured impedances (magnitude only) of CH1-CH4’s transformer-balanced low-Z inputs, on the “normal” impedance setting (using full primary coil), versus frequency. Data-point symbols show the measurements, and they are linked by smooth curves to represent each channel’s characteristic (CH1 and CH3 were indistinguishable).

By comparison, switching to “low-Z” (handle up; transformer primary tap used) while keeping the pad off (“normal”), drops measured input impedances at each given frequency by about 4-fold (curves not shown in Figure 15 to avoid clutter). Measured impedance at 1 KHz is then 48 Ω for CH1 and CH3, 470 Ω for CH2, and 98 Ω for CH4. This 4-fold impedance drop is consistent with halving the number of turns in the primary winding (confirming the tap is a true center tap, i.e., with equal turns on either side). In an ideal transformer, the impedance in the secondary circuit

is reflected on the primary according to the square of the turns ratio, so if primary turns decreases by ½, input impedance becomes $(½)^2$ = ¼ of the full-coil value.

Engaging the pad switches for the CH1-CH4 balanced inputs causes their impedance to range between 1130-1260 Ω regardless of the channel, frequency, and low-Z switch setting. Series resistors R1 and R3 in the pad networks (see Figure 8) dominate input impedance in this case, as explained in the following section (see Figure 16, Model B, Equation 2).



[EQ. 1] INPUT IMPEDANCE:

$$Z_{IN} = Z_{LOAD}$$

[EQ. 2] INPUT IMPEDANCE:

$$Z_{IN} = R1 + R3 + \left[\frac{1}{\frac{1}{Z_{LOAD}} + \frac{1}{R2}}\right]$$

[EQ. 3] LOSS (dB), DUE TO Z_{SOURCE}:

$$\text{LOSS} = 20 \log \frac{Z_{LOAD}}{Z_{SOURCE} + Z_{LOAD}}$$

[EQ. 4] LOSS (dB), DUE TO Z_{SOURCE} AND PAD:

$$\text{LOSS} = 20 \log \frac{\left[\frac{1}{\frac{1}{Z_{LOAD}} + \frac{1}{R2}}\right]}{Z_{SOURCE} + R1 + R3 + \left[\frac{1}{\frac{1}{Z_{LOAD}} + \frac{1}{R2}}\right]}$$

[EQ. 5] DECREASE IN AMPLITUDE WHEN PAD IS SWITCHED ON (dB):

$$\begin{bmatrix}\text{PERCEIVED LOSS} \\ \text{DUE TO PAD (dB)}\end{bmatrix} = \begin{bmatrix}\text{LOSS DUE TO } Z_{SOURCE} \text{ AND PAD (dB)} \\ \text{[MODEL B, EQ. 4]}\end{bmatrix} - \begin{bmatrix}\text{LOSS (dB), DUE TO } Z_{SOURCE} \\ \text{[MODEL A, EQ. 3]}\end{bmatrix}$$

Figure 16. Modeling impedances and signal losses for CH1-CH4's low-Z balanced input pads. Model A (left column) applies when pad is off, and Model B (right column) is with pad engaged; both models are recruited for Equation 5 (at bottom); see text.

5.3. Channels 1-4: Modeling Balanced Pads. Figure 16 models source coupling to CH1-CH4's balanced inputs with the low-Z pad off (Model A) or on (Model B). Importantly, these simple models ignore reactive components of the source or load impedances; therefore, please

consider their accuracy as limited to the audio mid-band (around 1 KHz). Each diagram shows source impedance (Z_{SOURCE}) as two equal resistors, one in series with each wire of the balanced line. The circled waveform icon is a hypothetical AC generator (voltage source) with zero impedance, so the individual resistors for the balanced Z_{SOURCE} are in series and simply sum together. (Z_{SOURCE} could be diagrammed as a single resistor without altering the math.)

With the pad turned off (Model A in Figure 16), Equation 1 simply says that a channel's input impedance (Z_{IN}) equals Z_{LOAD}. Z_{LOAD} is the impedance of a channel's transformer-balanced input circuit (including the 220-Ω shunt resistor in the case of CH1 and CH3), and is symbolized by a resistor in these models (remember that model accuracy is restricted to the audio mid-band). The voltage loss (in dB) across Z_{SOURCE} in this case is given by Equation 3, in which Z_{SOURCE} and Z_{LOAD} form a voltage divider. Note how matched source and load impedances result in a 6-dB voltage loss, as mentioned in Section 5.2.

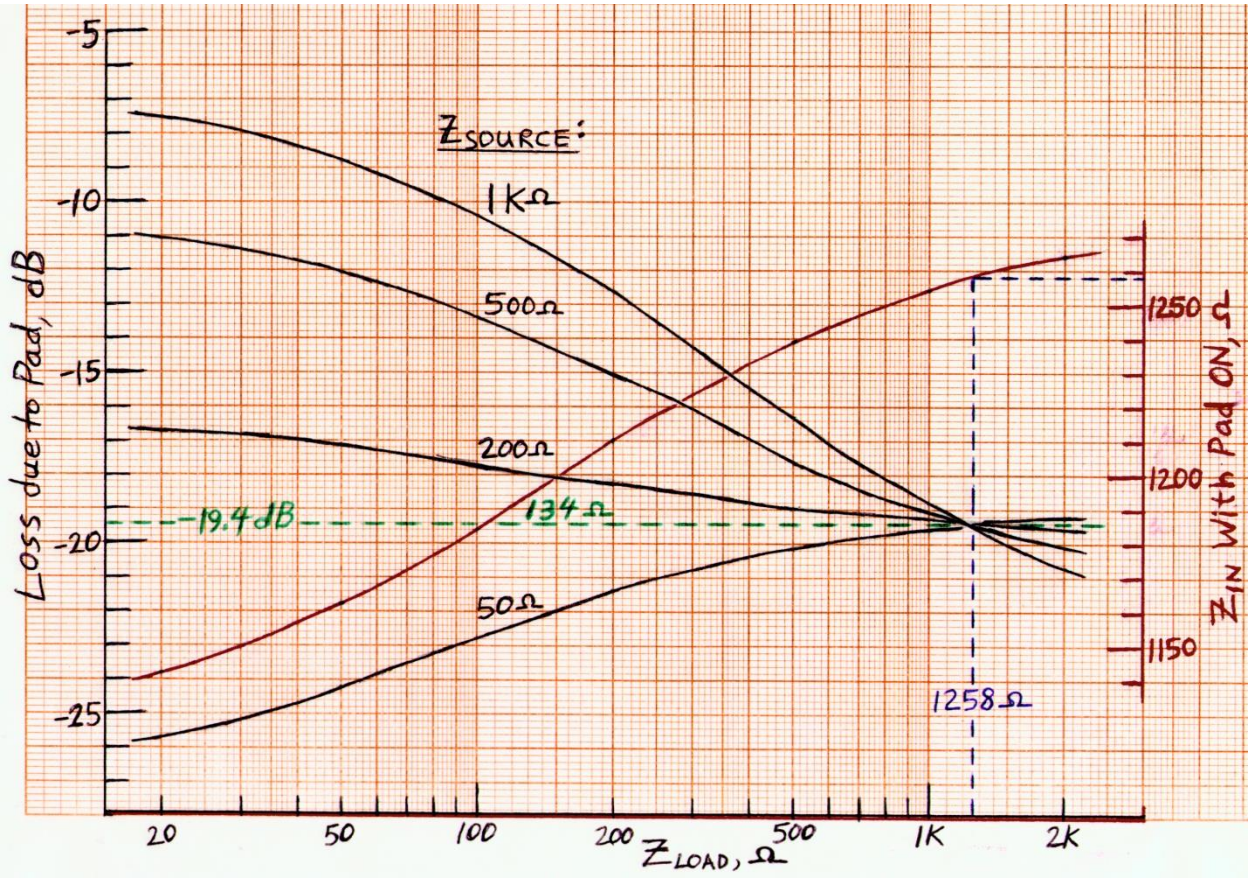


Figure 17. Performance of balanced input pads used in CH1-CH4 of modified Altec 1567A, as predicted from models shown in Figure 16. The horizontal axis (Z_{LOAD}) is the input impedance when pad is switched off. The input impedance with a pad engaged is shown by the red curve (plotted against red vertical scale at right), which is the solution to Equation 2 in Figure 16. The perceived loss when a pad is engaged is given by the black curves (using black scale at left), which emerge from Equation 5 in Figure 16; loss depends on Z_{LOAD} except when Z_{SOURCE} = 134 Ω (dashed green curve). Engaging a pad changes the input impedance except when Z_{LOAD} is 1258 Ω (dashed blue index lines).

The pad shown in Model B (Figure 16) has the same component numbers (R1-R3) and values as used in CH1-CH4 of the modified unit (see Figure 8). Equation 2 says that a channel's balanced input impedance (Z_{IN}) when the pad is switched on is the sum of series resistors R1 and R3, plus the parallel combination of Z_{LOAD} and R2. This equation generates the red curve in Figure 17, which shows that the pad converts a two-decade Z_{LOAD} range (20 Ω to 2 KΩ) into less than a ten-percent range of Z_{IN}. Engaging the pad increases Z_{IN} when Z_{LOAD} is less than 1258 Ω (emphasized by dashed blue index lines in Figure 17).

Equation 4 in Figure 16 expresses the combined loss across the pad and source impedances when a pad is engaged. But this is not the difference one hears when switching the pad on, because there is already some "hidden" loss across Z_{SOURCE} with the pad off (given by Equation 3). To model the perceived loss when a pad is turned on, the solution of Equation 3 must be subtracted from that of Equation 4, as stated in Equation 5 (Figure 16, bottom). The black curves in Figure 17 represent Equation 5 solved for a range of Z_{SOURCE} values; the dashed green horizontal line is the special case of Z_{SOURCE} = 134 Ω, which yields a perceived pad loss of 19.4 dB at all Z_{LOAD} values. All loss curves converge on -19.4 dB when Z_{LOAD} is 1258 Ω, which is the Z_{LOAD} value at which Z_{IN} does not change when a pad is turned on (dashed blue lines).

Summarizing Figures 16 and 17, the effect of CH1-CH4's low-Z balanced pads depends on both source and load impedances. (If you infer that I designed these pads targeting 20 dB loss at Z_{LOAD}s in the 1-2 KΩ range, you would be correct; I lowered the impedance of in CH1 and CH3 by adding the 220-Ω shunts after installing the pads.) Also, recalling the way each channel's Z_{LOAD} depends on frequency (Figure 15), bandwidth restriction due to pad engagement will occur under many conditions (even though I again caution you against using my scalar impedance measurements and models for accurate bandwidth prediction). Unfortunately, no single pad design can give consistent performance when source and load impedances vary (the latter depending on channel, low-Z switch setting, and frequency). Yet, these channels have high gain (see Section 5.5) and there will be cases when pads are needed, such as using efficient mics on loud sources. I advise users to be aware of the imperfect pads and keep a keen ear on their sonic results, so that the compromises necessary for pad design do not sneak into the artistic product. If a high-output mic has its own built-in pad, try that pad first.

5.4. Channels 1, 2, and 5: Impedance and Compatibility of Unbalanced Inputs. In each channel equipped with a high-Z unbalanced input jack, the signal is applied to the top (clockwise-most) terminal of a 1-MΩ pot (via a coupling capacitor in the case of CH5). The wiper of this "pre-triode attenuator" pot hooks directly to the grid of a 12AX7 triode (see Figures 10 and 14). The resistive component of the unbalanced input's impedance remains near 1 MΩ regardless of the attenuator and high-Z pad settings. However, inter-electrode capacitances in the triode, including the gain-dependent "Miller Effect," reduce input impedance in a frequency-dependent manner; total capacitance is about 104 pF at the gain used. When the pre-triode attenuator and pad are in their normal positions (full clockwise and off, respectively; see Section 5.13), this drops the magnitude of the unbalanced input impedance from near 1 MΩ at 20 Hz, to 840 KΩ at 1 KHz, to about 80 KΩ at 20 KHz, according to the model shown in Figure 18. One needs to bear this in mind when using extremely high-Z sources such as piezoelectric pickups.

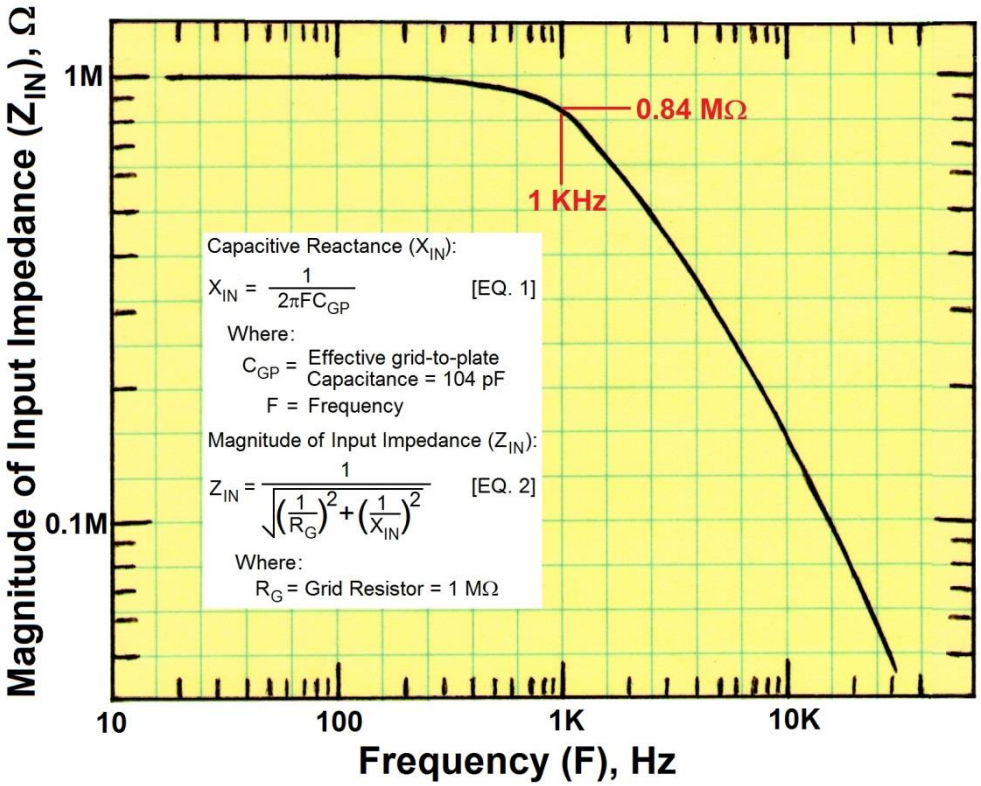


Figure 18. Input impedance model for the unbalanced, high-Z inputs of CH1, CH2, and CH5 with pre-triode attenuator full clockwise and high-Z pad off. Equation 1 shows how the triode's capacitive reactance depends on frequency, and Equation 2 places that reactance in parallel with the 1-MΩ grid resistor. The curve is the solution to Equation 2. Red index lines indicate performance at 1 KHz.

With minimum-length patch cables recommended, the unbalanced inputs are compatible with the outputs of the modified unit's other channels, as well as most external low-level ground-referenced sources. External sources should either share a good common ground with the modified Altec unit or receive their ground reference via their output patch (as in an effects pedal or an electric guitar or bass). "Line-level" sources on the nominal -10 dBV standard should usually be compatible, but if peaks cause unwanted distortion, decrease output level at the source if possible; avoid the temptation to turn down the high-Z input's pre-triode attenuator knob unless killing some high end is specifically desired (see Section 5.13).

One final remark about the unbalanced inputs: note in the schematics (Figures 10 and 14) that these inputs are direct-coupled in CH1 and CH2, but AC-coupled via capacitor C1 in CH5. While direct connection in CH1 and CH2 keeps input coupling like that of a stock Altec 1567A, be aware that DC offset at those unbalanced inputs will cause distortion proportional to the amount of offset. The CH5 input is immune to DC offset. Significant offsets are not present in the modified unit's own outputs, and they should be rare in external sources.

5.5. Channels 1-4: Gain. In each of these channels, voltage gain occurs at three stages: (1) input transformer, (2) triode circuit, and (3) output driver. For individual stages and whole

channels (pads off, attenuators and faders full clockwise), gain measurements were made at 1.0 KHz by comparing input and output voltages read on the Hewlett-Packard 331A Distortion Analyzer's RMS voltmeter. Input signals (from the B&K Precision 3011B generator) were low-amplitude to insure a low-distortion output. Measuring actual amplitudes across inputs makes gain figures independent of source and input impedances. Output readings used normal (or defined) load conditions; for example, an isolated triode circuit was loaded by the voltmeter's 1-MΩ input Z shunted by a 330 KΩ resistor for a load equivalent to a 250- KΩ fader pot. Such a measurement equals the voltage output of the channel's high-Z unbalanced output working into an open circuit (note that load presented by the output driver circuit's very high impedance is negligible). Channel gains were equal within ±0.75 dB and the average result is given here.

With the "normal" input transformer impedance setting, in which the input couples across the full primary winding, the transformers' voltage gain is 25 dB. On the "low-Z" setting, which uses the center tap, transformer gain is 30.3 dB. This increase of 5.3 dB is less than the expected 6.0 dB when halving the primary-to-secondary turns ratio of an ideal transformer. Being real, the Altec model 4722 input transformers are expected to fall short of ideal performance, probably due mostly to loss in the DC resistance of the secondary winding. However, I should note that my tests on the "low-Z" setting were not as extensive (replicated) as on the "normal" setting; some measurement error is also possible.

The triode circuits deliver 35 dB gain when loaded by 250 KΩ, the fader pot resistance in these channels. Note that gain will decrease according to the load on a channel's high-Z unbalanced output (see Section 5.8). Based on a 12AX7A characteristic chart published by RCA in 1960, a predicted gain for this circuit is 30.3 dB. The better-than-predicted gain could be a function of the excellent vintage Telefunken 12AX7s used, or 12AX7As may be slightly different, or my prediction may have some error because I used classic graphical (load-line) techniques. I tested my favorite vintage RCA 12AX7 specimen (mid-1960's, black plates) in a breadboard version of the circuit and obtained 35.4 dB gain. It's a good bet that most high-quality 12AX7s will provide such high gain.

Each channel's THAT1646 balanced line driver IC adds 5.5 dB gain when driving 600 Ω loads (and nearly 6.0 dB into 10 KΩ or more), whether operating in differential or single-ended output mode. This is a result of the way these chips balance the output to emulate a transformer, and is described further in Section 5.9. The total gain available in CH1-CH4 is summarized in the table at right. It is based on

GAIN (dB)		INPUT		
		UNBALANCED (CH1 and CH2 only)	BALANCED	
			Normal Impedance (Full Primary Coil)	Low Impedance (Primary Coil Tap)
OUTPUT	UNBALANCED (Into ∞ Ω Load)	35	60	65.3
	BALANCED (Into 600 Ω Load)	40.5	65.5	70.8

the individual stage gains just discussed, and confirmed by direct measurement for the most common configurations (such as full-coil balanced input to low-Z balanced output: 65.5 dB ±0.75 dB on all channels).

5.6. Channels 1-4: Cross-Talk. Cross-talk between channels results from (1) coupling through their common power supply and (2) electromagnetic coupling through space. As noted in Sections 4.3 and 4.4, star-grounding and an individual B+ decoupling network for each channel improves the original Altec 1567A design, helping to limit the first source of cross-talk. The second source is addressed by shielding the most sensitive (high-impedance, low-input-amplitude, high-gain) circuits and maximizing their distance from outputs.

For pairs of channels with their faders set full clockwise (and no pre-triode attenuation for CH1 and CH2), I evaluated cross-talk by comparing amplitudes at the balanced outputs. On one channel, the balanced input was fed a 1-KHz test signal with amplitude sufficient for maximum un-clipped output into a 600-Ω load (called 0 dB for this test). With the other channel's input open, its measured output level was expressed in dB relative to that of the active channel. I could detect cross-talk only between pairs of channels that share a twin-triode (i.e., CH1/CH2 and CH3/CH4), suggesting electromagnetic coupling within and near tubes as the dominant path. The worst pair was CH1 bleeding into CH2, which read -45 dB; most of this susceptibility appears related to the relatively high impedance of CH2's balanced input (see Section 5.2) reflected to the grid circuit, because simply engaging CH2's input pad (which shunts the transformer primary with a 150-ohm resistor) reduced cross-talk to -64 dB. Cross-talk for other 12AX7-sharing pairs tested as follows: CH2 into CH1: -72 dB; CH3 into CH4: -78 dB; CH4 into CH3: -74 dB. While not specifically measured, cross-talk increased with increasing frequency.

When cascading channels to get distortion effects due to excessive gain, be alert to the likelihood of feedback. Cross-talk can close a positive feedback loop by coupling a portion of the cascade's output back to the input, causing oscillation when phases correlate properly. Thus, when adding up enough gain to cause severe distortion, always start with your monitors turned down very low until you are sure that the set-up is stable. Ear-splitting high-frequency oscillation cannot be avoided if you are going to explore all of the distortion possibilities this experimental equipment offers, so monitor very softly any time you attempt to increase distortion in cascaded channels. You can easily damage your ears (and those of others nearby) and/or monitors with this gear, so I cannot stress enough the caution you must exercise.

5.7. Channels 1-4: Frequency Response. The modified Altec 1567A has many control settings and I/O options, some of which affect frequency response. These include frequency-dependent impedance and the imperfect pads of transformer-coupled inputs as mentioned in Sections 5.2 and 5.3. The high-Z pads and pre-triode attenuators of CH1 and CH2 have profound effects that will be described in Section 5.13. Even the channel fader settings subtly affect frequency response as I will summarize shortly. The response curves in Figure 19 were measured under conditions that give nearly the best (flattest and widest) response possible in CH1 and CH2 using the balanced inputs (CH3 and CH4 data are similar to that of CH1 and are omitted to avoid clutter). Notice that CH1 (solid curve) has slightly greater bandwidth than CH2 (dashed curve); the presence of a 220-Ω shunt resistor across the primary of CH1's input transformer is

the only difference between these two channels. This may help explain why Altec included a similar resistor in the original design (see Sections 4.6 and 5.2).

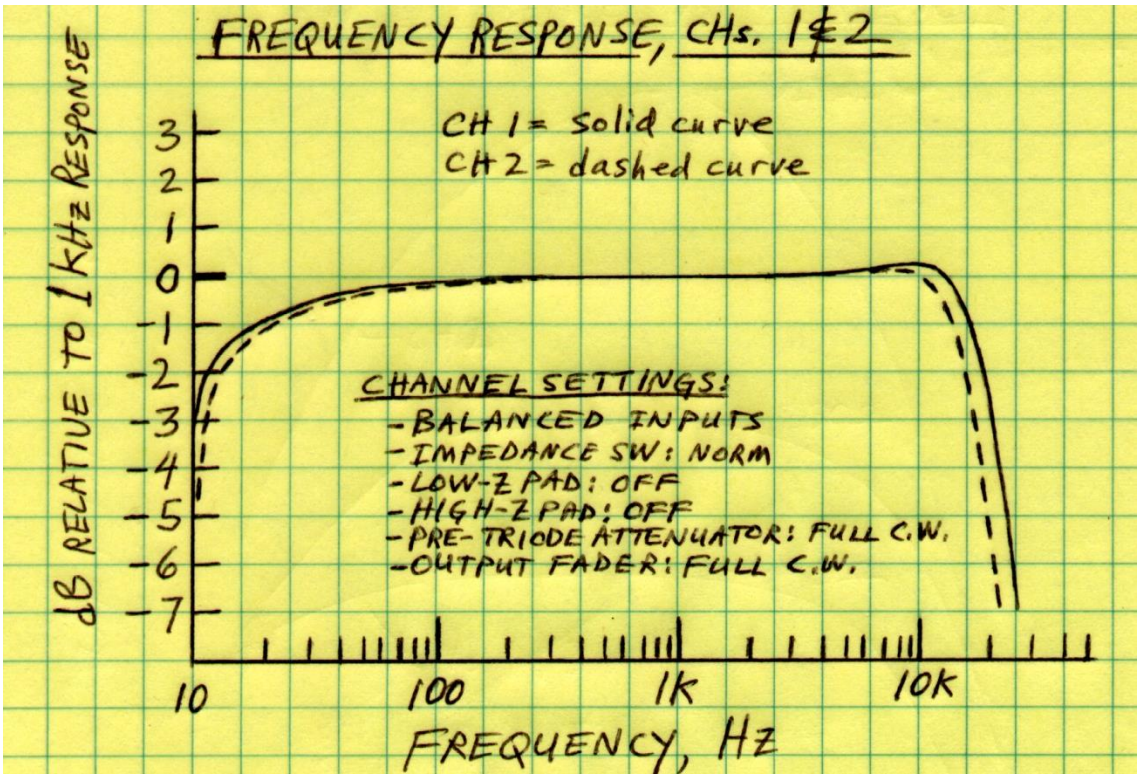


Figure 19. Measured frequency response of CH1 (solid curve) and CH2 (dashed curve) relative to response at 1 KHz = 0dB, under conditions listed in the drawing. Sine-wave input was from a function generator with 50-Ω source impedance. Balanced channel outputs were terminated with 600 Ω in single-ended mode, and measured with RMS voltmeter in the Hewlett-Packard 331A instrument. Results for CH3 and CH4 were the same as that of CH1.

The channel fader pots present a resistive load to the triode outputs (see Sections 5.5 and 5.8). Additionally, stray capacitance in the shielded cable connecting the pot's wiper to the output driver board (and high-Z outputs) apparently causes wiper position-dependent impedance changes that affect frequency response slightly. At lower settings ("32 dB" to about "12 dB" as painted on the vintage panel), there is a broad response peak centered on about 12-18 KHz (depending on channel) that rises above the 1-KHz reference response by no more than +4 dB. This high-frequency emphasis flattens out as fader settings increase beyond "18 dB" to "12 dB."

This frequency response data uses CH1-CH4's solid-state buffered outputs; the output driver circuits impart virtually no load on the fader pot wipers. Results will vary when the high-Z unbalanced outputs are used, depending on the impedance of the device driven and the length of the shielded patch cable used. The next section discusses the impedance of the unbalanced outputs and how it depends on fader setting.

5.8. Channels1-4: Unbalanced Output Impedance and Applications. At the triode operating point used in CH1-CH4, the dynamic plate resistance (R_P) is about 87.5 KΩ; for these common-cathode circuits with cathode bypass, source impedance at the plate is R_P in parallel with the 220-KΩ plate load resistor, or 62.6 KΩ. With a 250-KΩ fader pot set full clockwise (0 dB attenuation), output impedance for the unbalanced line is 50 KΩ (62.6 KΩ in parallel with 250 KΩ). As the fader is rotated counter-clockwise, visualize the wiper as dividing the 250-KΩ pot resistor into "top" (clockwise-most) and "bottom" portions. Output impedance becomes the "bottom" resistance in parallel with the series combination of the "top" resistance and 62.6 KΩ. Attenuation relative to the full clockwise setting (in dB) is 20 times the logarithm of the fraction: "bottom" resistance divided by 250 KΩ. This math yields the following table:

Fader Attenuation, dB	Unbalanced Output Z, Ω
0 (full CW)	50K
-4.1	78K (maximum Z_{OUT})
-8	68K
-14	42K
-20	23K
-28	9.7K
-34	4.9K
-40	2.5K

I did not determine how accurately the faders are calibrated on the vintage panel, but -40dB (bottom row of table) is marked close to full counter-clockwise (note that Altec omitted the "-" or minus signs in labeling the faders). In any case, most fader settings require that the unbalanced outputs feed fairly high-impedance inputs to avoid significant loss. For example, a standard unbalanced 10-KΩ "line level" input would cause an additional -6 dB loss at about the "28" fader setting and a 15.6 dB loss at full clockwise. On the other hand, patching these outputs to the modified unit's own high-Z inputs (provided on CH1, CH2, and CH5) will cause little loss, as will patches to "instrument-level" inputs on guitar amps, pedals, et cetera.

Strictly speaking, CH1-CH4 can't be considered "all-tube and –transformer" signal paths when using just the high-Z unbalanced outputs; the diode clamps (D1 and D2 in Figure 12) protecting the solid-state output stages (see Section 4.8) also affect the high-Z outputs at high amplitudes. Hard-clipping limits the output swing to ±18V (that's 12.7 V_{RMS} or 22.1 dBV; also see Figure 26). This level is so high that practical situations where clipping occurs should be most infrequent. If necessary, the clip alert indicators can be set to indicate diode clipping by adapting the calibration procedure described in Section 5.10.

5.9. Channels 1-4: Balanced Output Characteristics. The THAT1646 balanced output driver ICs used in CH1-CH4 act like output transformers, except in at least five ways: (1) Across the audio band, output impedance is essentially independent of frequency. (2) A broad range of input impedances can be driven directly; a load resistor shunt is not required when feeding higher-Z inputs (e.g., 10-KΩ "line-level" inputs). (3) When saturated, the distortion mode is hard-clipping. (4) Ground-referencing one output leg for single-ended operation nearly halves the

clipping threshold (unless current-limited; see below). (5) As set up in CH1-CH4, the driver chips may be damaged if hooked to mic preamp inputs with phantom power active, so this condition needs to be carefully avoided.

As a direct approach to determine balanced output impedances (Z_{OUT}), I measured output voltage with (E_{LOAD}) and without ($E_{NO\ LOAD}$; typically used 1 V_{RMS}) various known load resistors (R_{LOAD}) across the output (between XLR pins 2 and 3, with pin 3 grounded), then solved the equation: $Z_{OUT} = [R_{LOAD}(E_{NO\ LOAD} – E_{LOAD})]/E_{LOAD}$. (Note that the 1-MΩ impedance of the H-P 331A's voltmeter that I used negligibly affects results for low-Z outputs in this method.)

Output impedance of the THAT1646s as deployed in CH1-CH4 tested 57 Ω at 1 KHz (within THAT's 60-Ω maximum spec, but 14 % greater than the nominal 50-Ω). These solid-state devices emulate transformers by using feedback to control the two complementary outputs legs. As a by-product of this approach, the driver effectively doubles the input voltage (6 dB gain) at high load impedance, or nearly doubles it (5.5 dB gain) working into a 600 Ω load (the difference is due mainly to voltage drop across the output impedance). This is because the voltage swing of each complementary output leg matches that of the input, so across a floating (differential) load, the difference voltage is twice that of the input. In single-ended mode, where one output needs to be grounded (exactly as a transformer-balanced output is used in single-ended mode), feedback automatically forces the opposite output to nearly double the input voltage. Using a single output leg without grounding the opposite leg is not recommended, since this lets common-mode noise—normally squelched by feedback—appear at the output; this increases the chip's normal noise by over 40 dB. Normal driver noise is described further in Section 5.14.

Similarity to transformer-balanced outputs breaks down at the clip thresholds, not just because of the distortion type (hard clipping), but because the clip threshold is about 5 dB lower for single-ended mode than differential mode (at 600 Ω load). That's because an individual output cannot swing beyond a limit set by the power supply voltages, and feedback confers gain on one output leg when the other is grounded, as discussed above. By measuring output amplitudes across different load resistors while carefully observing the waveforms on an oscilloscope, I gathered the data shown in Figure 20; this clearly shows the amplitude difference between the differential and single-ended clip thresholds (black and red curves, respectively), at least for load resistances of 300 Ω or more.

As shown in Figure 20, load resistance has a relatively small effect on clip threshold when it is greater than about 300 Ω in single-ended mode or 600 Ω in differential mode. In this load range, the output voltage swing is limited by the driver's power supply voltage; the slight threshold decrease as load resistance decreases is consistent with voltage drop across the driver's 57-Ω impedance. As load resistance decreases beyond 300 Ω (single-ended) or 600 Ω (differential), clip thresholds decrease more steeply. In this range, the slope suggests clipping is caused by an instantaneous current limit of about 57 mA (approaching the chip's rated short-circuit current of 70 mA).

The two highlighted data points in Figure 20 represent line driver clipping into the balanced inputs of CH1 and CH2 at 1 KHz, rather than into dummy load resistors. These loads are the channels' respective input impedances at 1 KHz (from Figure 15). Relevant when experimenting

with input transformer saturation, one of CH1-CH4's balanced outputs can deliver 8 dB higher amplitude to CH2's balanced input than it can to CH1 before clipping at 1 KHz. This is discussed further in Section 5.12. In general, the dependence of CH1-CH4's buffered output clip threshold on load impedance requires careful consideration when adjusting the clip alert indicators, as addressed in the next section.

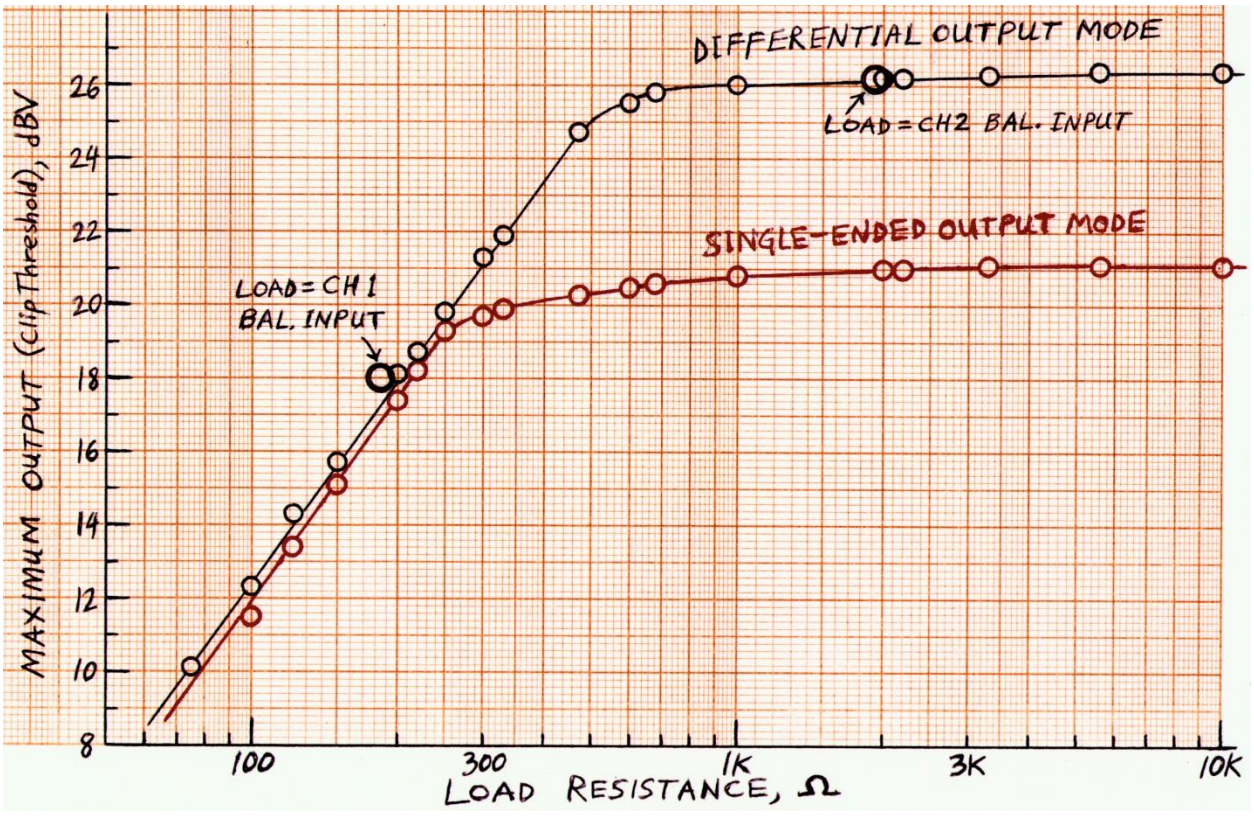


Figure 20. Threshold for clipping of 1-KHz sine waveform at CH4's balanced output operating in differential (black) or single-ended (red) mode; CH1-CH3 performance should be identical. Thresholds judged using oscilloscope, and RMS outputs measured on H-P 331A's voltmeter. Small data-points represent measurements using resistors as loads, and the two large black data-points represent the transformer-balanced input of CH1 or CH2 as load, and are plotted using their impedance at 1 KHz (see Figure 15).

5.10. Channels 1-4: Understanding and Adjusting Clip Alerts. As described in Section 4.9, the clip alert circuits do not work by directly detecting clipping at the balanced line driver outputs. Instead, they are threshold detectors monitoring the driver chips' inputs. Since the output clipping threshold depends on load and whether the output mode is differential or single-ended (see previous section), meaningful use of the clip alerts requires their adjustment for the specific output conditions used.

Prior to shipping the modified Altec 1567A, I set CH1-CH4's clip alert thresholds to indicate clipping into floating (balanced or differential-mode) 600-Ω load resistors. Referring to Figure 20, one can see how this setting gives reasonably accurate (about ±1 dB) performance for balanced loads 470 Ω and greater (i.e., for clipping mainly due to the balanced driver's voltage limit).

However, using single-ended mode with the same loads would let severe clipping go undetected by the alert circuit. Staying in balanced mode but dropping the load impedance to about 200 Ω would have the same result. Conversely, adjusting the clip LEDs to trigger on such lower thresholds would give a false indication of clipping for high-impedance balanced loads.

Given the high-amplitude capability of these drivers, clipping is not likely to be a concern except when over-driving a following stage for distortion effects, such as with input transformer saturation experiments. CH1 and CH2 are equipped for such experiments since they have pre-triode attenuators in their input transformers' secondary circuits. As shown in Figure 15, CH2's balanced input impedance exceeds 470 Ω between about 33 Hz and 7.5 KHz. Within this band, clipping due to driver saturation is accurately indicated (±1 dB) by the associated clip alert LED as presently adjusted (for 600-Ω loads). As explained in Section 5.12, transformer saturation in CH2 can severely distort bass frequencies at amplitudes well below the driver's clip threshold.

If required, CH1-CH4's clip alerts are easily adjusted if a function generator, oscilloscope with 10X probe(s), and a small screw-driver are available. A spare female XLR connector (with shell removed), clip leads, and various resistors can be used to test different loads on the outputs; connect such dummy load resistors between pins 2 and 3. Or, when observing clipping into the input of another channel or device, the shell of the patch cable's female XLR end should be loosened and slid back for access to the terminals while plugged in. In either case, for differential (balanced) output mode, oscilloscope probe(s) can hang on either XLR pin 2 or 3 (or both, if multiple 'scope channels are available) and their grounds clipped to pin 1. Do not ground a probe to an output leg (pin 2 or 3) unless single-ended mode is specifically desired (in which case it is best to tie pin 1 directly to either pin 2 or 3 anyway).

Set your function generator for a low amplitude and patch its output into an input of the channel to be adjusted (a pad may be required if the generator's minimum output exceeds -40 dBV or 10 mV_{RMS}). A 1-KHz sine waveform makes a good test signal, but sometimes triangular waveforms are better for observing nascent clipping on the 'scope. With the channel fader (and pre-triode attenuator for CH1 and CH2) at full clockwise, slowly increase the generator's output amplitude until clipping is seen on the oscilloscope display, then back it off just enough so that no clipping is seen. The output buffer is now operating at its clip threshold. Using a fine-tipped slotted screwdriver, turn that channel's clip alert trim-pot clockwise if the nearby red LED is already on, or counter-clockwise if it is off; continue turning until the LED changes state. Fine-adjust the trim-pot so that the LED turns on when the output's clip threshold is just exceeded as judged using the 'scope.

5.11. Channels 1-4: Types of Distortion. There are at least five types of harmonic distortion (distortion affecting a waveform's harmonic composition) offered by CH1-CH4. While it may seem odd to discuss distortion as a performance feature, remember that one goal of this modified Altec 1567A is to permit controlled amounts of distortion for musical effect. All but the first of these could be useful: (1) Hard-clipping by the solid state output buffers or their input-protection diode clamps (discussed above). (2) "Routine" even-order harmonic distortion by the triode stages. (3) Soft-clipping by saturated triode stages. (4) "Routine" low-level distortion from

magnetic hysteresis in the input transformers. (5) Distortion due to input transformer core saturation. I will discuss the latter four in turn.

Even-order harmonic distortion accounts for much of the sought-after “warmth” that good tube gear offers. It is always present (hence, “routine”) using CH1-CH4, because each channel’s triode gain stage lacks feedback. Such a stage has a non-linear transfer characteristic—incremental changes along the input waveform’s voltage axis do not map to exactly proportional changes in the output waveform. For example, the output voltage swing caused by a change of instantaneous input voltage from (say) +100 to +150 mV is greater than that caused by a -100 to -150 mV input change. This asymmetry adds even-order harmonics, accounting for much of the “musicality” of triodes. With faders full clockwise and input amplitude just sufficient to give the maximum un-clipped buffered output of 20.5 dBV (single-ended mode into 600 Ω; 1 KHz), my Hewlett-Packard 331A instrument measured total harmonic distortion (THD) for CH1-CH4 at about 0.5 percent. Greater THD is available by increasing the input amplitude while compensating with the fader (and less obtains by decreasing the input amplitude). In Section 5.14, the right-hand graphs in Figures 25 and 26 show how distortion depends on output amplitude for a triode circuit like those used in CH1-CH4.

Theoretically, since common-cathode stages invert the signal, the second of two identical cascaded channels could either remove or increase “routine” harmonic distortion produced by the first, depending on the second channel’s polarity switch setting (the first channel’s fader attenuation should equal its gain for this to possibly work). Note that CH1-CH4’s polarity switches affect only their balanced inputs, and that the balanced output connection already compensates for inversion by the triode stages as described in Section 4.8. So with a balanced patch between channels, the second channel’s “normal” polarity setting would tend to compound harmonic distortion, while it’s “invert” setting should subtract it. This is but one of many experiments with sonic “character” worth a try on the modified Altec 1567A.

At the triode operating points for CH1-CH4, triode saturation (over-driving) occurs when instantaneous positive grid voltage approaches the cathode bias voltage (near +1 V). Current through the triode is then maximal (further positive voltage swing at the grid cannot cause current to increase). Voltage drop across the plate load resistor is maximum so the output signal flattens out (soft-clips) on the negative (bottom) side of the waveform. An inverted example is in the bottom oscilloscope trace on the right-hand side of Figure 21 (inverted because the polarity-compensated, buffered output fed the ‘scope). Clipping is “soft” because thresholds are imposed relatively gradually with vacuum triodes. Indeed, the point at which “routine” harmonic distortion gives way to “soft-clipping” is not distinct, but it occurs as distortion reaches 3 to 5 percent (see Figures 25 and 26 in Section 5.14). Such limiting is less harsh-sounding than hard clipping. Sufficient over-drive of CH1-CH4 also soft-clips negative input peaks as triode current gradually transitions toward “cut-off” (for instantaneous grid voltages less than about -3.5 V). However, at such high amplitudes, clipping on the positive side is so severe that the results may be quite harsh (even “soft” clipping has its practical limits). Soft-clipping in CH1-CH4 is always asymmetrical, favoring even-order harmonics.

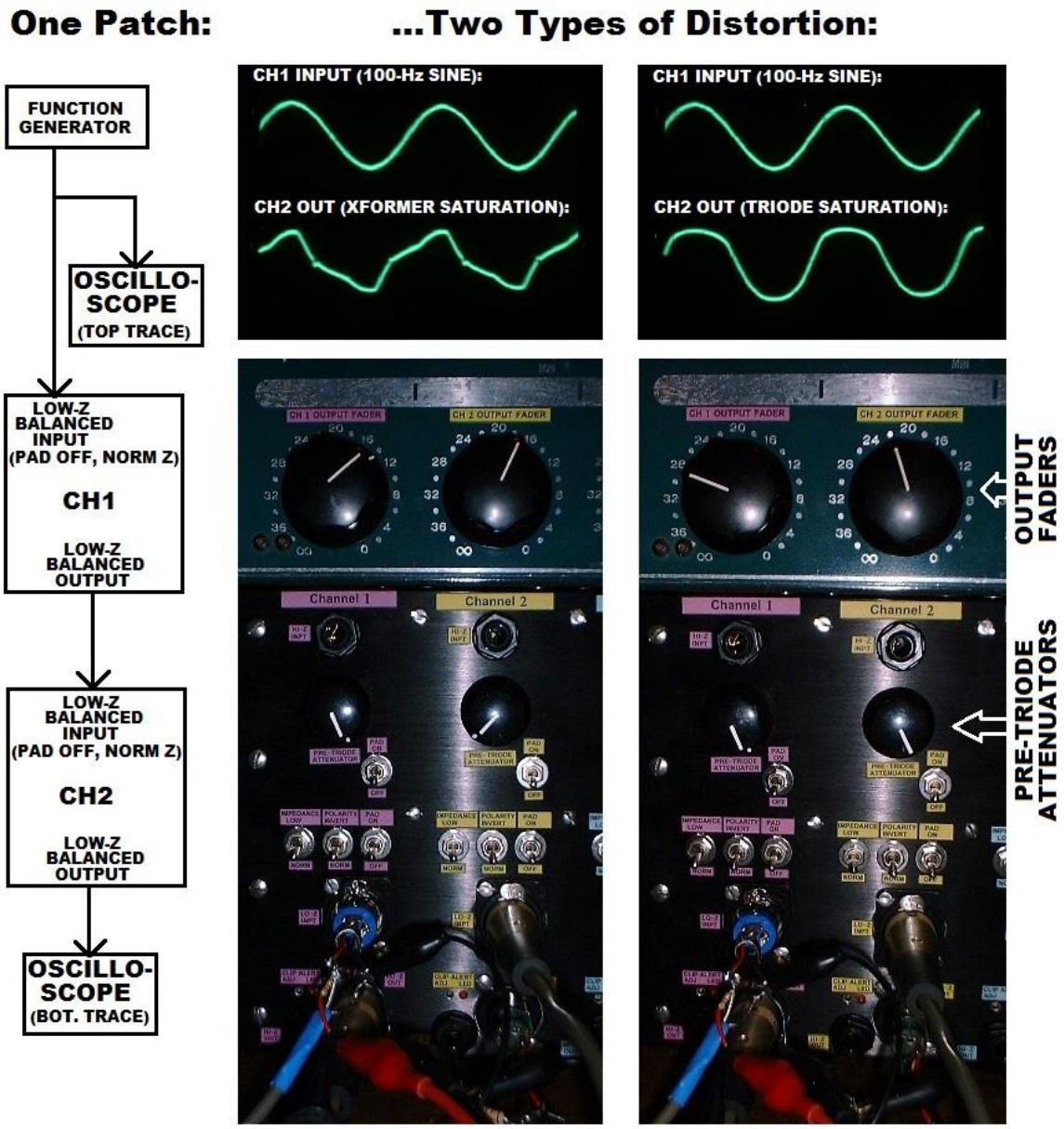


Figure 21. Two examples of waveform distortion available by cascading channels of the modified Altec 1567A. In this patch (block diagram at far left), CH1's balanced output feeds CH2's balanced input. CH1's input signal is a 100-Hz sine wave (top oscilloscope traces). Images of the front panel control settings used to demonstrate input transformer saturation (left photo) or triode saturation (right photo) are below the corresponding oscilloscopic results (bottom traces are waveforms at CH2's output).

All real audio transformers routinely distort signals due to a magnetic "memory effect" of the iron-alloy core, called hysteresis. Essentially, hysteresis makes the device's transfer characteristic have two separate (but close) parallel curves, one for each direction of current. Normally (in the absence of a DC offset or core magnetization) waveform distortion is symmetrical and odd-order harmonics are favored. The relative amount of distortion can be large for very low amplitude signals (quite loosely analogous to digital audio's quantization

distortion), and it affects low frequencies the most. Perhaps transformer distortion accounts for some of the “vintage character” audio engineers seek in their tone. An excellent resource for transformer topics is Chapter 11 (“Audio Transformers” by Bill Whitlock) in “Handbook for Sound Engineers, Third Edition,” Glen M. Ballou, Editor, 2002, Focal Press (a PDF file of that chapter is available at http://www.jensen-transformers.com/an/Audio%20Transformers%20Chapter.pdf.

Only CH1 and CH2 are set up for practical experiments to hear input transformer saturation. A sufficiently high amplitude signal on the primary causes magnetic flux density in the core to reach a maximum (saturate). With practicable input amplitudes, only the lower end of the audio band can be targeted, as explained in the next section. Alternating magnetic flux density in the core matches the phase of the current in the windings, which lags the voltage by 90 degrees. So flux maxima occur when instantaneous voltage is crossing zero in a sine waveform. When flux maxima are “clipped” due to core saturation, the transformer’s output voltage waveform is upset mostly in the zero-crossing excursions, not clipped at their maxima (similar to cross-over distortion in a poorly-biased class B power amp). Unlike triode saturation, the distortion is symmetrical and adds odd-order harmonics. A rather extreme example is shown in Figure 21 (bottom trace in left-hand ‘scope display). Core saturation is covered further in the next section, and Section 5.13 deals with performance of the necessary pre-triode attenuators.

A significant aspect of Figure 21 is that two very different types of distortion are obtained using the same two-channel cascade; the only differences are the fader and pre-triode attenuator settings. Note CH2’s pre-triode attenuator is set near full counter-clockwise for transformer saturation. (Incidentally, the two photos of the panel happen to be from slightly different angles, so they form an ersatz stereo pair. Try viewing them with crossed eyes for a 3-D-like effect.)

5.12. Channels 1 and 2: Input Transformer Saturation Threshold. A context for experimenting with transformer saturation to distort audio programs is to understand the susceptible frequency range and the amplitudes needed. With their pre-triode attenuators in the extreme counter-clockwise range to compensate for excessive input levels, I fed CH1 or CH2’s transformers from the balanced output of CH4. The CH4 input was a sine waveform of various frequencies, with amplitude too low for triode saturation; its output was thus “clean” up to the solid-state driver clip threshold. Amplitude at the balanced input of CH1 or CH2 was measured using the H-P 331A’s RMS voltmeter, while an oscilloscope compared CH1 or CH2’s output waveform to that of CH4’s input. At distortion thresholds, it’s easy to distinguish driver clipping from transformer saturation because they manifest at maximum excursions versus zero-crossings, respectively.

Theoretically, maintaining a given magnetic flux amplitude (including the saturation threshold) as frequency doubles requires doubling the input voltage. This is because flux amplitude is proportional to current in the windings, which, due to their inductive reactance, require double the voltage to maintain constant current as frequency doubles. Thus the saturation threshold should rise at 6 dB per octave. Shown in Figure 22, my actual results approximate this prediction, but the observed slope is closer to about 7 dB per octave. It is unclear whether the difference is due to: (1) the transformer is not ideal; (2) the test signal is not a pure sine

waveform, having been amplified in CH4's triode stage; (3) measurement error; or (4) a combination of these.

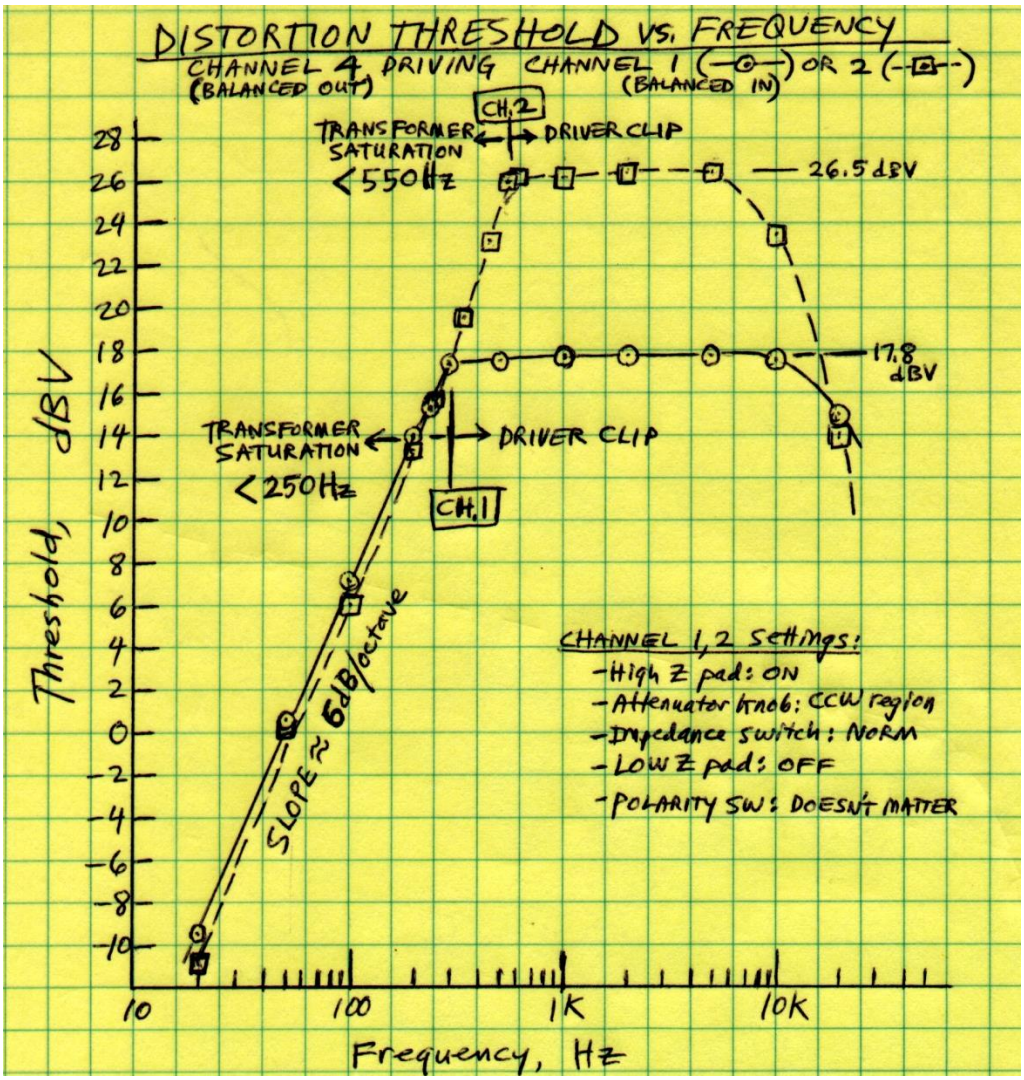


Figure 22. Distortion threshold (and type) versus frequency with CH4's balanced output driving the balanced input of CH1 (circular data-points/solid curve) or CH2 (square data-points/dashed curve). CH1 and CH2 settings are listed in drawing. At each sine-waveform frequency tested, distortion thresholds were judged using an oscilloscope on the output of CH1 or CH2; RMS voltage at CH4's output was then measured. The type of distortion was also noted: When due to transformer saturation, distortion appears at the waveform's zero-crossing phases. Distinctly, clipping at the output driver manifests at peak phases (both positive and negative).

In any case, Figure 22 shows that transformer saturation gives way to driver clipping at 250 Hz for CH1 and 550 Hz for CH2. This is consistent with the differential-mode output driver clipping characteristic shown in Figure 20, and CH1 and CH2's input impedance versus frequency shown in Figure 15. The 220-Ω primary-shunt resistor in CH1 lowers the input impedance by providing a pathway for current to bypass the transformer's primary. The driver cannot source

sufficient current to saturate the transformer at frequencies greater than 250 Hz and still heat up the shunt resistor. The only difference with CH2 is that CH2 lacks a shunt resistor. This buys 8 dB more mid-range headroom (note in Figure 20 that the driver is current-limited when clipping into CH1 at 1 KHz). It also widens the band susceptible to transformer saturation (by slightly more than one octave), when driven by one of the modified unit's other hybrid channels.

Therefore Figure 22 suggests that CH2 is a better candidate than CH1 for hearing what transformer saturation offers as a distortion effect. Only the low frequencies of an audio program are subject to this effect. Perhaps it could be tried on a kick drum or bass guitar to see if this type of distortion does anything aesthetically useful in the context of a mix.

5.13. Channels 1, 2 and 5: Pre-Triode Attenuator/High-Z Pad Applications and Effect on Bandwidth. In CH1 and CH2, the pre-triode attenuators and pads allow compensation for the very high amplitudes at the input transformer's secondary when experimenting with transformer saturation. In CH2, for example, when the full-coil primary is driven at 25 dBV in the mid-range band, the 25-dB voltage step-up by the transformer makes the amplitude across the secondary 50 dBV, a shocking 316 V_{RMS}! To safely avoid triode saturation, at least 58 dB of attenuation is required before applying this signal to the grid. Additional specialized experimental uses for these attenuators may include directly interfacing speaker-level outputs from power amps to the unbalanced inputs of CH1, CH2, or CH5. However, for all routine applications, attenuators should be set full clockwise, and the high-Z pads turned off; input amplitudes causing unwanted triode saturation should be turned down at their source whenever possible. With high-output microphones on loud sources, the low-Z (not high-Z) pads should be used (and if the mic has its own pad switch, it should be tried first).

The reason pre-triode attenuation should be avoided is that it can cause significant bandwidth loss depending on the setting. The problem is stray capacitance associated with the attenuator networks and their I/O lines. Rather than attempting a formal model, I will give my observations first and then remark on how stray capacitances may explain them. Using CH1 (and assuming results for CH2 and CH5 would be similar) at different pre-triode attenuator and high-Z pad settings, I measured response relative to that of 1 KHz for frequencies of 1 KHz and greater. With source Z = 50 Ω, the sine wave generator fed CH1's balanced input, which was set for full primary coil (nominal 150 Ω) and low-Z pad off. With fader full clockwise, amplitude was measured at the low-Z output, which was used in single-ended mode terminated with 600 Ω.

Frequency response with the high-Z pad turned off is portrayed on the right-hand side of Figure 23; the pre-triode attenuator settings that were examined are diagrammed below the curves. To help sort out the curves, those in red represent clockwise-range settings until the setting with maximum high-frequency roll-off is reached at 3:00 ("three o'clock"), and the black curves are more counter-clockwise settings. While high-frequency loss is present for most settings, some high-frequency emphasis and bandwidth extension appears by 9:00 as the pot is turned counter-clockwise; flattest response in this counter-clockwise region must be between 9:00 and 10:00. With the high-Z pad switched on (left-hand side of Figure 23), bandwidth is severely limited when the pre-triode attenuator is fully clockwise, but improves for all other settings. Flattest response is somewhere between 9:00 and 10:30. Compared to the results with the

high-Z pad turned off, the high-frequency emphasis at 9:00 is greater (3.4-dB emphasis at 27.5 KHz with pad on versus 1.2-dB emphasis at 26.1 KHz with pad off).

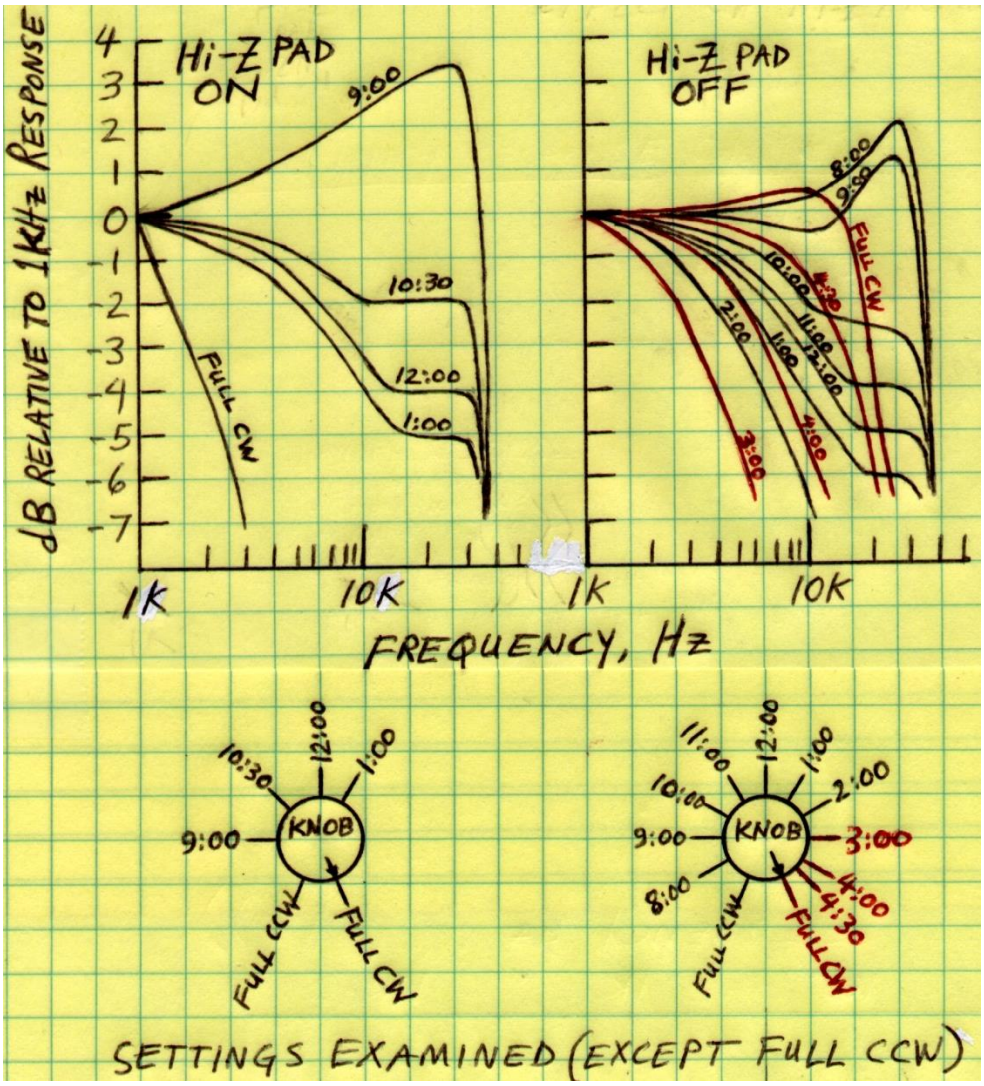


Figure 23. Effect of CH1's pre-triode attenuator and high-Z pad setting on measured frequency response relative to response at 1 KHz = 0 dB. Same conditions and method as used in Figure 19, except frequencies <1 KHz were not examined, and various pre-triode attenuator settings plus both high-Z pad settings were evaluated. Response curves are at top; diagrams at bottom show the pre-triode attenuator settings examined when high-Z pad was on (left) or off (right). In the latter case, red settings correspond to red response curves, which highlight the approach to the bandwidth minimum at about 3:00 (three o'clock) as the attenuator is turned counter-clockwise.

Another way to show the data is Figure 24, where the bandwidth is directly plotted against pre-triode attenuator setting. Here, bandwidth is expressed as the frequency at which response (compared to that at 1 KHz, which defines 0 dB) crosses below -1 dB. As the attenuator is turned clockwise (left to right along horizontal axis), the bandwidth difference due to high-Z pad setting is small until 3:00, beyond which bandwidth continues to degrade if the pad is on, but

starts to recover if the pad is off. Fortunately for transformer saturation experiments (see Section 5.12), sufficient attenuation requires settings in the counter-clockwise end of the range (although the high-frequency emphasis there may be unwanted and require EQ). For virtually all other applications and routine use, these attenuators should be left fully clockwise and the high-Z pads switched off.

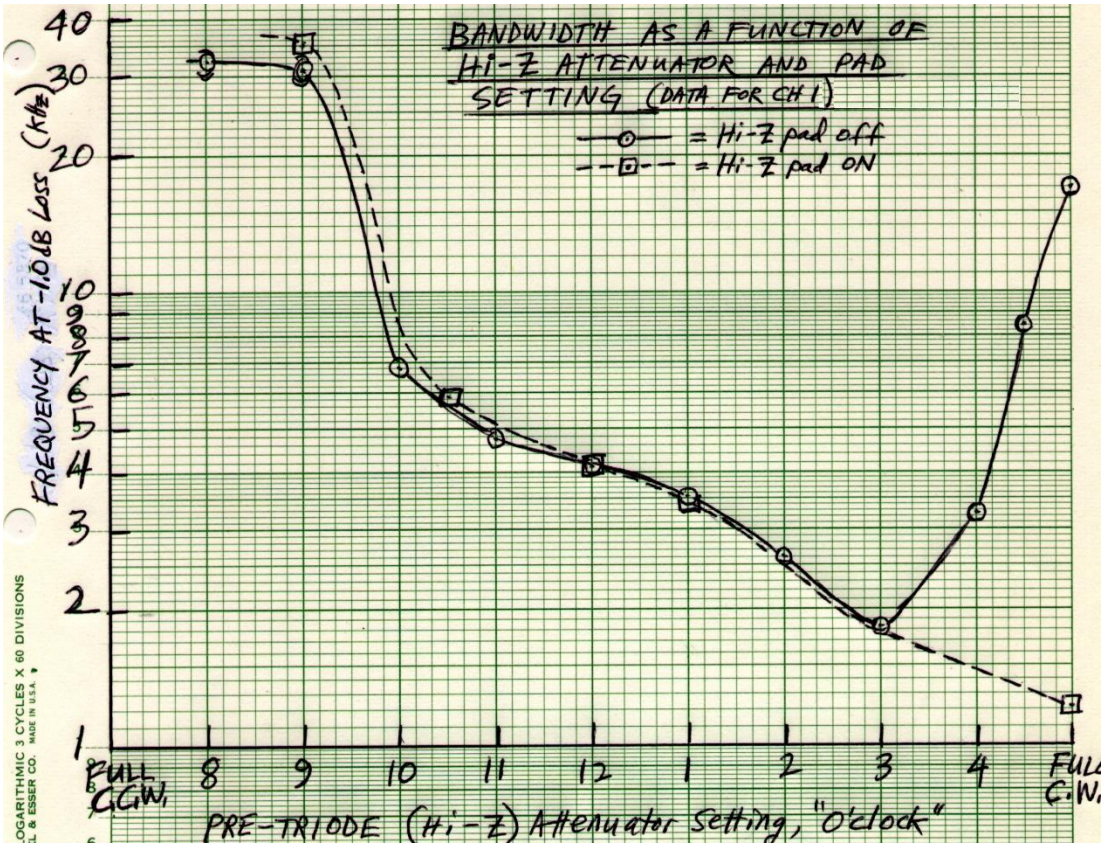


Figure 24. Effect of CH1's pre-triode attenuator setting on bandwidth, with high-Z pad off (circular data-points/solid curve) or on (square data-points/dashed curve). This is derived from Figure 23's dataset; here, each curve shows the frequency at which response is -1.0 dB relative to the response at 1 KHz, as a function of attenuator setting.

Since my tests used CH1's balanced input, the source impedance seen by the pre-triode attenuator network was probably on the order of 50 KΩ (the nominal secondary impedance of the Altec 4722 transformer). Expect additional high-frequency loss in the attenuator's clockwise range (with pad off) when source impedances greater than 50 KΩ are inserted at CH1, CH2, or CH5's unbalanced inputs; the severity of this, and the need for short patch cables, will increase as the source Z increases. At least this is suggested by my informal analysis, which follows.

Here is how stray capacitance may explain the observed results (you may skip this paragraph if you like your analyses air-tight). The most problematic stray capacitance apparently acts between the pot's wiper and ground, like hooking a capacitor in parallel with the pot's "bottom" (counter-clockwise) resistance. This provides a low-Z path to ground (bypass) for high frequencies. With the high-Z pad switched off, the relative effect of this capacitor is greatest

when the resistances “above” (including the source impedance in series with the pot) and “below” the wiper’s position are equal. This is the wiper position giving the attenuator network its greatest output impedance (the parallel combination of resistances “above” and “below” the wiper). The relative effect of high-frequency bypass diminishes toward the extreme pot settings where impedance is lower. (This is a log-taper 1-MΩ pot, explaining why maximum impedance and minimum bandwidth is offset clockwise, to 3:00. Applied to the “top” of the pot, the source impedance [about 50-KΩ in this case] also helps shift the Z-max clockwise.) When the high-Z pad is switched on, output impedance of the attenuator network continuously increases as the pot is turned clockwise; relative high-frequency bypass by the stray capacitance is maximum at full clockwise in that case. A second, smaller stray capacitance might explain the high-frequency emphasis in the counter-clockwise range: this capacitance would be between the input and output of the entire attenuator network. This would let high frequencies escape attenuation, and its relative contribution would be inversely related to the amount of attenuation.

5.14. Channels 1-4: Noise. Subjectively, these channels are quiet for tube gain blocks and compare favorably to similar equipment I’ve listened to. Objectively, my noise measurements require a few assumptions, which I will make conservatively. For each channel with fader set either full counter-clockwise or full clockwise, I measured total noise using the Hewlett-Packard 331A’s voltmeter on XLR pin 2 of the buffered outputs in single-ended mode, loaded at 600 Ω. The balanced channel inputs were open (not terminated, but recall that CH1 and CH3 each have a built-in 220-Ω shunt resistor), all relay control switches were in normal (down) positions, and no pre-triode attenuation or high-Z pad was applied on CH1 and CH2. Readings differed by less than 1.5 dB between channels; I will present results from CH1 as typical.

With full counter-clockwise fader, channel noise comes only from the solid-state output stage; the raw measurement was 283 μV_{RMS}. The voltmeter’s own noise (its input shorted to ground) was 68 μV_{RMS}. (The H-P 331A’s voltmeter is “average-responding,” so all noise readings have been multiplied by 1.13 to get these “true” RMS figures.) Channel noise is unrelated to voltmeter noise, so subtracting the latter from the channel’s reading requires taking the square root of the difference between the two squared readings; this gives 275 μV_{RMS} as the corrected output stage noise voltage. But only the part that is in the audio band (20 Hz to 20 KHz) is relevant. The published small-signal bandwidth spec for the THAT1646 output driver chip is 10 MHz. A fair assumption is that its noise power is equal per frequency increment (i.e., “white” noise; equal power per Hz) extending out to 10 MHz. At this point, I need to digress about the frequency response of the voltmeter.

On the 1-mV range that I used, the H-P 331A voltmeter’s published bandwidth is 5 Hz to 3 MHz. This spec refers to the instrument’s flat (within ±0.45 dB) frequency response; signals beyond 3 MHz also deflect the meter. As H-P Application Note 206-1, dated October 1977 (available at http://www.hpmemory.org/an/pdf/an_206-1.pdf) puts it,

> “…it may be desired to measure the noise in an audio amplifier. In this case, we are interested in the noise only from 20 Hz to 20 KHz as this is the maximum range the ear can hear. This is easily done with the 3045A program [the spectrum analyzer discussed in Application Note], but without external filters the 331A will measure the noise out to several megahertz. As there could be significant noise beyond 20 KHz, the 331A might

> read considerably higher than the desired result. One must not depend on the roll-off of the audio amplifier to limit the noise as it often will not."

Indeed there's no reason to think the THAT1646 output driver's own noise is bandwidth-limited (until 10 MHz). Unfortunately I do not know the noise bandwidth of the H-P 331A more accurately than "out to several megahertz." So I will make an assumption. According to Walter G. Jung ("IC Op-Amp Cookbook" 3rd Ed., 1986, ISBN: 0-672-22453-4; p.44), if a system's bandwidth is limited by a single-pole low-pass filter (6 dB per octave roll-off) defined by -3 dB response at f_C, its noise bandwidth is $1.57f_C$. I will assume f_C = 3 MHz for the H-P 331A voltmeter, which is conservative because error is rated only ±5 percent (within ±0.45 dB) for 5 Hz to 3 MHz; the actual f_C must be higher. (However, its effective filter characteristic could be above first-order, with greater than 6 dB per octave roll-off. I'll assume it is first-order without evidence, other than to say it would have been harder for the instrument's designers to achieve flat response given a higher-order filter characteristic, nor should it be necessary to build in such response. If my function generator could go a couple octaves beyond its 2-MHz limit I would test, rather than assume.) Thus I conservatively call the voltmeter's noise bandwidth 4.7 MHz.

A 4.7-MHz bandwidth for the voltmeter allows expressing the output driver's 275-µV noise figure as 127 nV per root-Hz noise density, so the driver circuit's observed noise voltage is 17.9 μV_{RMS} (-95 dBV) in the 5 Hz to 20KHz band. The THAT1646's published output noise spec (balanced 600-Ω load, 0-Ω source, ±18 V supply, 22 Hz to 20 KHz bandwidth) is -101 dBu, which is 6.9 μV_{RMS} or -103 dBV. (The 5-Hz versus 22-Hz lower bandwidth boundary is insignificant. Also, the difference in source is effectively small: the modified Altec 1567A's unity-gain OPA2604 op amp driving the THAT1646 is rated very low noise at 10 nV per root-Hz translating to 1.5 μV_{RMS} for the audio band; uncorrelated with the driver's noise, this would account for only about 0.2 µV of the 17.9 µV figure.) The observed driver output noise is thus 8 dB larger than expected from THAT Corporation's specs. Much of this difference is likely due to a conservative estimate of voltmeter noise bandwidth, so actual driver noise is probably between -103 and -95 dBV (i.e., worse than in laboratory conditions but better than the figure I am reporting.)

With channel fader full clockwise (0 dB attenuation), the output noise reading was 565 μV_{RMS}, corrected to 561 μV_{RMS} due to voltmeter self-noise. I'll assume the triode gain circuit's noise lies entirely within the audio band, which is reasonable given Figure 19. As amplified by the output driver, triode circuit noise is then simply the square root of the difference between 561 μV_{RMS} squared and the broad-band driver noise (275 μV_{RMS}) squared, or 489 μV_{RMS} (-66.2 dBV). Accounting for 5.5-dB output driver gain when working into 600 Ω, triode circuit noise at the top of the fader is 260 μV_{RMS} (-71.7 dBV). Compared to output driver noise, triode noise is so large that the driver noise is not significant unless fader attenuation is between infinity and about -30 dB, as explained next.

Subtracting channel gain (see Section 5.5) from the -66.2 dBV output noise figure gives the equivalent input noise (EIN): referred to the balanced inputs, gain is 65.5 dB so EIN = -131.7 dBV; for the unbalanced inputs of CH1 and CH2, gain excludes the 25-dB input transformer step-up making EIN = -106.7 dBV there (this reasonably assumes that transformer noise is insignificant compared to the triode's). Triode stage noise decreases as a channel fader is

turned counter-clockwise, while output driver noise does not change. EIN therefore increases with fader attenuation as the relative contribution of driver noise increases (see table below). Of course, EIN is independent of input signal amplitude, but the signal-to-noise ratio (SNR) is not. SNR at a given output amplitude is easily found: using dB units, subtract the fader attenuation and the channel gain from the output amplitude to get the input signal amplitude, then subtract the EIN corresponding to that fader setting to find SNR. Along with EIN at different fader settings, the following table shows SNR when the balanced input signal is -45 dBV (this input signal yields the maximum un-clipped output amplitude [20.5 dBV; see section 5.9 and Figure 20] into 600 Ω in single-ended mode when the fader is full clockwise).

Fader Attenuation, dB	EIN, dBV[1]	SNR, dB[2]
0 (full CW)	-131.7	86.7
-10	-131.6	86.6
-20	-131.1	86.1
-30	-128.0	83.0
-40	-120.1	75.1

1. At balanced inputs; channel gain = 65.5 dB. Add 25 dBV for EIN at CH1/CH2 unbalanced inputs.
2. With input amplitude that yields 20.5 dBV S.E. output into 600 Ω at 0dB fader attenuation.

At any given fader setting, SNR decreases without limit as signal amplitude decreases. However, improving SNR by increasing the amplitude reaches a limit when additional triode distortion becomes intolerable and/or driver-stage clipping occurs. On the left-hand side of Figure 25, SNR is plotted against fader setting for five maximum (i.e., fader fully clockwise) output amplitudes in 10-dB increments beginning with 0.5 dBV. (Actual output amplitude is less than the curves' labeled values by the amount of fader attenuation.) Output driver clipping conditions for the single-ended and differential output modes are indicated by red shading. Solid curves embody my conservative driver noise evaluation, while the dashed curves are based on THAT Corporation's published figure; actual SNR is probably between these curves. Harmonic distortion due to the triode stage is shown in the right-hand graph, which is lined up with the SNR chart to compensate for the driver's gain (5.5 dB). Figure 25 is designed to show the trade-off between SNR and triode distortion. For example, 90-dB SNR requires at least one percent harmonic distortion; 100-dB SNR requires at least four percent, and so on.

Similarly, for CH1-CH4's unbalanced outputs, SNR and its relationship to harmonic distortion is given in Figure 26. This chart assumes a very high load impedance connected to these outputs. Clipping due to the diode clamps protecting the solid-state output drivers (see Sections 4.8 and 5.8) is indicated in red. For any given triode stage output amplitude, the fader setting has little effect on SNR. The very slight decrease in SNR at counter-clockwise settings is predicted from the thermal (Johnson) noise of the stage's output impedance (see table in Section 5.8 for output impedance at the fader's wiper).

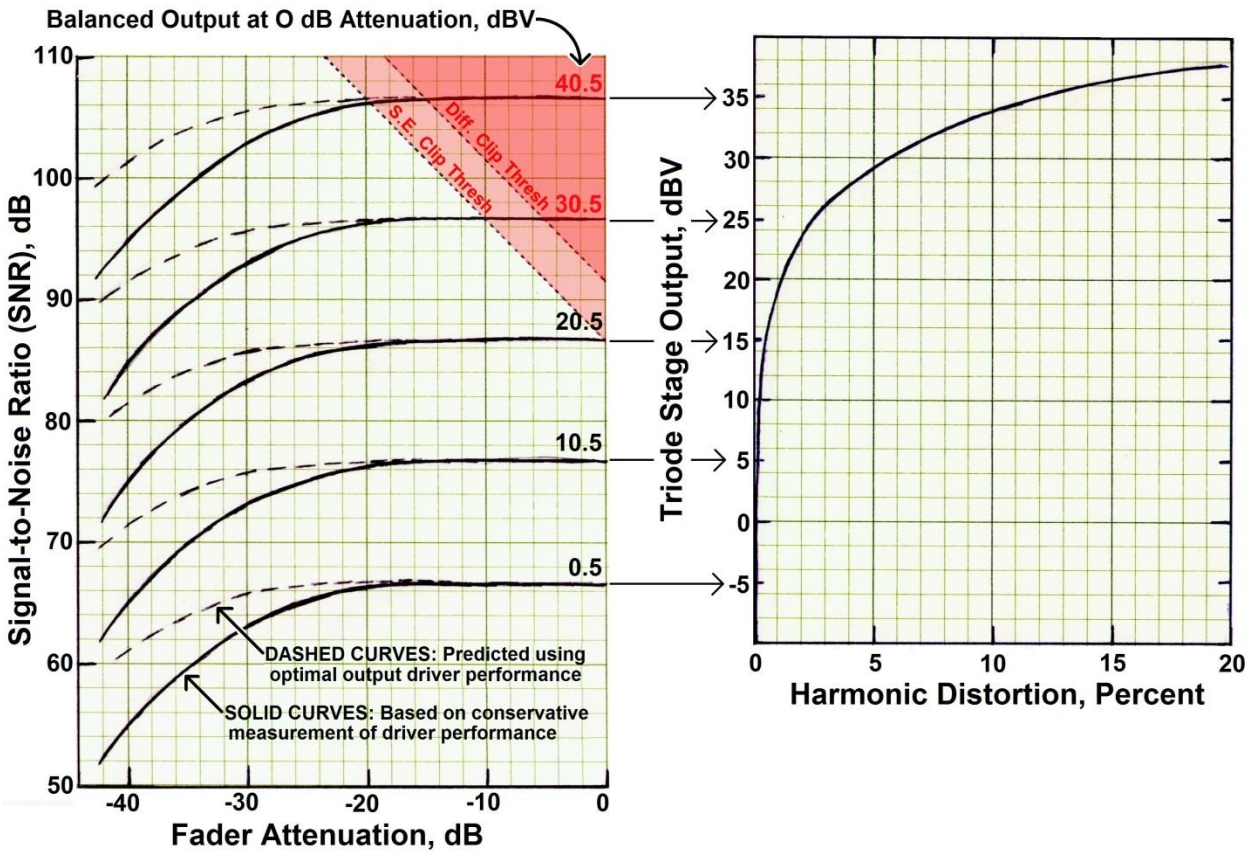


Figure 25. Left chart: Signal-to-noise ratio (SNR) at CH1-CH4's balanced outputs, terminated at 600 Ω, as a function of fader attenuation. Labeled output values for each pair of curves (solid and dashed) is output amplitude with full-clockwise fader (0 dB attenuation); add the fader attenuation to the labeled output value for actual output level at any fader setting. Solid curves use measured triode-stage and output driver noise (the latter being conservative); dashed curves use measured triode-stage noise and the published THAT1646 IC's noise figure (see text). Red-shaded area indicates clipping conditions for the line driver in differential mode (dark shading only) or single-ended mode (both light and dark shading). **Right Chart:** Percent harmonic distortion due to the triode stage, as a function of that stage's output amplitude. This chart is aligned vertically with the SNR chart so that SNR curves at full-clockwise fader align with the triode circuit output amplitude, as emphasized by the arrows (the 5.5-dB vertical offset equals the output driver's gain; see Sections 5.5 and 5.9). Distortion was measured on an isolated, breadboard version of CH1-CH4's triode circuit at 1 KHz using the H-P 331A distortion analyzer. The B&K Precision 3011B function generator that was used for test signals throughout this project delivered 0.6 % distortion with its 1-KHz sine waveform output; therefore, 0.6 % was subtracted from all raw distortion readings before stating distortion percentages in this report, including this chart. **Example:** Using -30 dB fader attenuation, say the RMS signal amplitude across a 600-Ω load is 0.5 dBV; this means the output would be 30.5 dBV at full clockwise fader, if the output driver stage could achieve that without clipping, which it can't. Nevertheless, the SNR curves labeled 30.5 dBV apply in this case. Thus, the SNR is between about 93 dB and 95.5 dB (depending on the driver noise evaluation used); the triode stage output is 25 dBV, and harmonic distortion is about 2.5 percent.

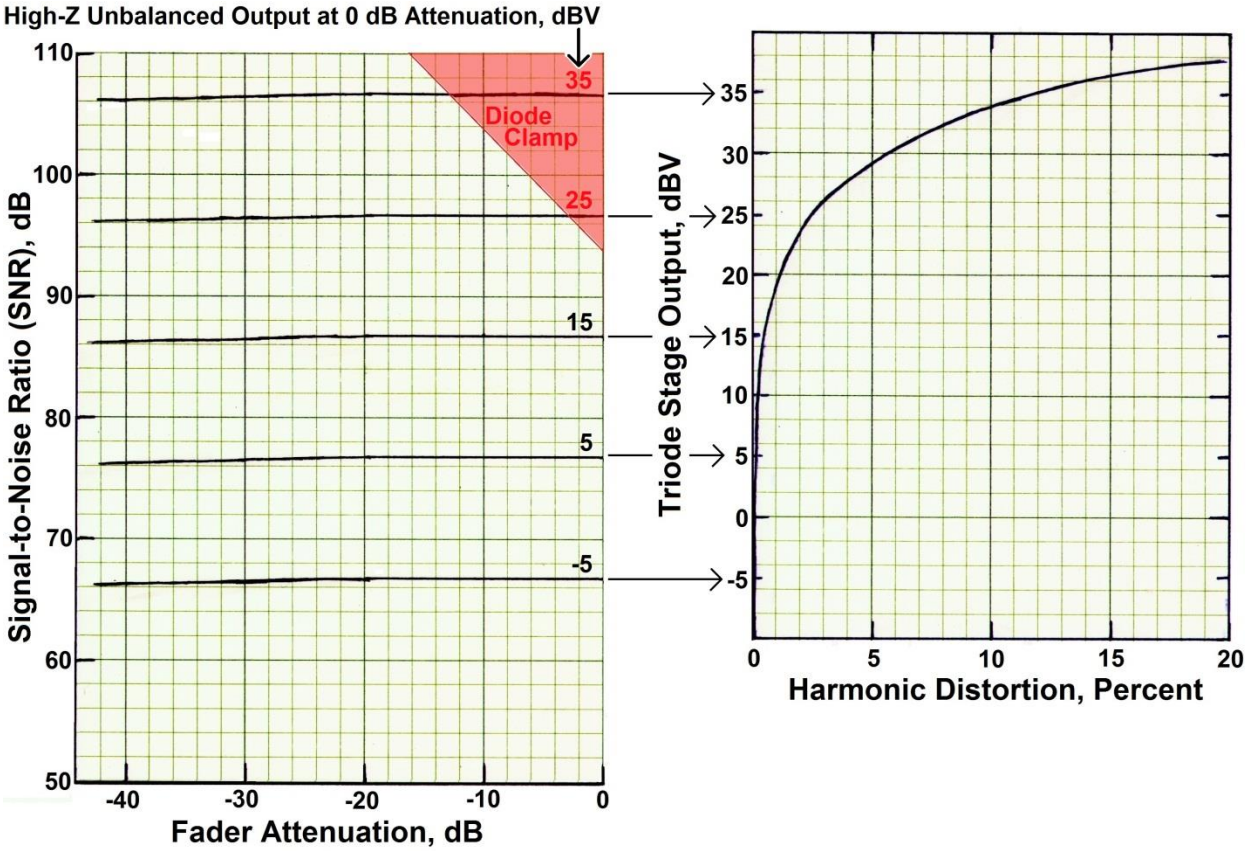


Figure 26. Left chart: Signal-to-noise ratio (SNR) at CH1-CH4's unbalanced, high-impedance outputs, as a function of fader setting (which makes little difference in this case; see text). Assumes very high load impedance (≥ 10 MΩ, say) on these outputs. Each curve corresponds to conditions giving the labeled RMS output amplitude at full-clockwise fader (0 dB attenuation); add fader attenuation value for actual output amplitude. Red shading indicates conditions causing clipping by diode clamps that protect the solid-state line driver circuits (see Sections 4.8 and 5.8) **Right chart:** Percent harmonic distortion versus triode-stage output (same as shown in Figure 25; see that figure's legend). Alignment to the left chart places full-clockwise-fader output amplitudes in register with the right chart's triode stage output scale, as indicated by arrows.

5.15. Channel 5: Applications and Input Characteristics. Since CH5 is based on the original master channel, one obvious application is using it to complete a vintage Altec 1567A signal path: simply patch the high-Z output of one of the other channels into CH5's input. With some adaptor-assembly effort, one might even reconstitute the original four-into-one mixer function: branch the 1/4-inch plug that feeds CH5's input to four 1/4-inch female jacks via 330K resistors (see R18-R21 in the original schematic shown in Appendix). Use a small shielded break-out box and minimum-length patch cables to the CH5 input and the CH1-CH4 outputs to be mixed.

Since CH5 lacks a low-Z balanced input that could be used with a microphone, a common application may be as an input channel for instruments. Thus, it can act as an active direct interface (DI box), complete with gain and tone controls. Of course, one (or both) of CH5's outputs can drive CH1-CH4 for extra gain and triode saturation distortion effects. But unlike

CH1-CH4, it may not cleanly deliver sufficient output amplitudes for input transformer saturation experiments. Also, since CH5 is noisier than CH1-CH4 (see Sections 5.14 and 5.19), expect quieter results when CH5 is the second, rather than the first, channel of a two-channel cascade.

The input impedance of CH5 is similar to that of CH1 and CH2 when the latters' unbalanced inputs are used (see Figure 18 and Section 5.4). This is not surprising, given their similar high-Z input circuitry (identical except CH5 is AC-coupled; see sections 4.7 and 4.10). As with CH1 and CH2, CH5's pre-triode attenuator should be kept fully clockwise and the high-Z pad switched off for most applications, except for rare cases when extreme attenuation or a high-frequency roll-off is desired (see Section 5.13 for characteristics of the high-Z attenuator/pad switch). If first-stage triode saturation is not wanted and input amplitude is too high, level should be turned down at the source, not at the pre-triode attenuator. Maximum high-Z input peaks falling below the triode saturation threshold are about 1.7 dB lower for CH5 than for CH1-CH4 (in Figures 14 and 10, note that cathode bias is 0.82 V for CH5's first triode [V3A] versus about 1.0 V for CH1-CH4).

5.16. Channel 5: Keeping Track of Knobs. Control of amplitude and tone in CH5's signal path is quite versatile but involves five knobs. The one on the auxiliary panel (pre-triode attenuator) should usually be fully clockwise as already discussed, so I will focus on the other four, which are on the vintage panel. The bass and treble knobs control a tone network (or "stack," as it was sometimes called) driven by input-stage triode V3A (see Figure 14). Neutral settings of these controls are near their 12-o'clock positions, but not exactly; I marked the positions that gave the best 1-KHz square waveform (on oscilloscope) for the overall channel, and hence represent an approximately flat frequency response (this test was done with the feedback control set at the vintage design level, marked near 12-o'clock). Marked knob positions can be seen in Figure 1.

Directly following the tone network along CH5's signal path is the knob I call the "channel fader" or the channel's "output fader" (formerly the original design's master volume control). Normally, one thinks of an output fader as the final amplitude control point of a channel, after any gain control. But this does not apply to CH5 as modified: the variable feedback knob affects gain downstream of the fader—in the channel's two-stage output driver. This is despite the counter-intuitive placement of the feedback knob to the left of the fader (everyone knows signals flow from left to right). Even though one can easily turn the output fader clockwise to saturate the first triode in the output driver (V3B in Figure 14) for distortion effects, there is no downstream attenuator pot to compensate for increased level. Generally, the feedback control does not affect a sufficient range of gain to be used for that purpose, and works by a different principle. Instead, the feedback control has its own interesting effects on distortion and tone, and interacts with the fader in complex ways to allow a range of voicing best explored by trial-and-error. Feedback control performance is described more specifically in the following section.

5.17. Channel 5: Variable Feedback Feature. The vintage Altec 1567A used a fixed negative feedback loop in the master channel, but as described in Section 4.10, the modified version recruits the un-used passive input channel's fader to serve as CH5's variable feedback control. This adds significant sonic variety to CH5, because changing the feedback level not only changes the output driver's gain, but affects its linearity: increasing feedback linearizes

performance, thus reducing the “routine” even-order harmonic distortion that triode stages add (Section 5.11 introduces how non-linearity causes distortion). Lowering feedback by turning the control clockwise increases gain while letting the channel’s second and third triodes (V3B and V4 in Figure 14) add more distortion to the signal. Reducing feedback also increases output impedance (discussed further in Section 5.20) and reduces bandwidth. In short, changing the negative feedback level affects all major aspects of amplifier performance.

Near its 12-o’clock position, I marked the feedback knob setting resulting in the same feedback loop resistance as that of the vintage Altec 1567A, called the “design feedback level.” Using moderate fader settings and a low-amplitude 1-KHz sine waveform input, I measured the changes in channel gain and distortion caused by different feedback settings. The total gain range available between the feedback knob’s extreme positions is about 11 dB. Relative to gain at the design feedback level, full counter-clockwise subtracts about 4 dB and full clockwise adds about 7 dB. (Importantly, the full clockwise setting is minimum *available* feedback, not zero feedback or open-loop.) The following average increases in total harmonic distortion were measured as feedback was decreased: from full counter-clockwise to the design feedback level, 0.14 %; from the design feedback level to full clockwise, 0.40 %. These figures should be taken as approximate, since it would require more care than I used to isolate CH5’s interacting distortion sources in its three-triode signal path (my distortion tests sometimes changed the input amplitude and/or fader position as well as the feedback knob; see also Section 5.19). However, they suggest an effect whose magnitude is reasonably close to the theoretical one, in which percent distortion is halved for each 6-dB decrease in gain due to feedback.

Since the vintage unit was originally designed for a specific fixed feedback level, frequency response was probably carefully tweaked for that condition (C14 in the original schematic is a good candidate for such a tweak). Making feedback variable by simply swapping a variable resistor for a fixed feedback resistor is a somewhat crude technique. And the extra physical length added to the feedback loop (using shielded cables to/from the vintage panel-mounted control) invites effects of stray capacitance. Therefore I was not surprised to measure frequency response effects of varying the feedback. As the feedback knob is turned clockwise (increasing gain), a high-frequency emphasis (around 7 KHz and above) is present at low gain, decreasing in intensity until just clockwise of the design level mark; flattest frequency response is between 1- and 2-o’clock; beyond that setting, high frequencies roll off with increasing gain until, at full clockwise, response relative to 1 KHz is -1 dB at 6.7 KHz and -3 dB at 16.5 KHz. To summarize, mid-range feedback settings yield the flattest frequency response.

5.18. Channel 5: Gain and Bandwidth. Unless noted otherwise, all of the figures reported in this section were with variable feedback set at the design feedback level (see preceding section), the tone controls neutral, and the pre-triode attenuator full clockwise. The channel fader was full clockwise for gain measurements. At the balanced output, which was set for nominal 600 Ω (secondary windings linked in series) and terminated with a 600 Ω resistor, channel gain for a low-amplitude 1-KHz sine wave input measured 54.5 dB; with the balanced output still under load, gain at the unbalanced output measured 70.1 dB into a 1-MΩ load (or a calculated 70.9 dB if the balanced output were not loaded). The difference in channel gain seen at the balanced versus unbalanced outputs represents the voltage difference between the

secondary and primary sides of the output transformer. As detailed in Section 5.20, this combines the effect of a 4.90:1 voltage step-down ratio and resistive losses in the windings. For the nominal 150-Ω setting (secondary windings in parallel), the transformer step-down ratio was 9.94:1 after accounting for resistive losses; at that setting, channel gain at the balanced output measured 50.1 dB with a 600-Ω load, and was calculated as 48.5 dB for a 150-Ω load.

Unlike for CH1-CH4, I did not measure the gain of CH5's individual stages, just the overall channel gain as given above. However, I have made the following estimates: Since CH5's first triode stage is very similar to those of CH1-CH4, its gain is likely near 33 dB (after accounting for an estimated 2-dB loss due to loading by the tone control network [with its knobs in the "neutral" positions] and 500-KΩ fader pot). This suggests that post-fader gain (the V3B/V4 output driver stages) is 37.1 dB under the conditions where 70.1 dB overall pre-transformer channel gain was measured (above paragraph). The two post-fader triode stages are in a variable feedback loop, and the minimum feedback setting increases gain by 7 dB compared to the design feedback level (see Section 5.17); guessing again, if this minimum feedback setting causes -2 dB gain reduction compared with open-loop gain, then the estimated open-loop gain of the V3B/V4 stages would be 37.1 + 7 + 2 = 46.1 dB. While modeling the output impedance of the V4 stage (Section 5.20), I also calculated its open-loop voltage gain as 9.5 dB. Therefore the estimated open-loop gain of the V3B stage is 46.1 – 9.5 = 36.6, which is within the reasonable range for a 12AX7 triode stage, absent feedback (note also that this value is a focal-point for accumulated errors in this estimate-rich analysis).

As noted in the previous section, the flattest frequency response is obtained when the feedback knob is between 1- and 2-o'clock (i.e., 1:30), just clockwise of the design feedback level mark. At that setting, relative to 0 dB at 1 KHz, frequency response measured at the balanced output terminated by 600 Ω was: -3 dB at 102 Hz; -1 dB at 227 Hz; a gradual, barely significant peak reaching +0.25 dB centered at 10.6 KHz; -1 dB at 20.8 KHz. With the balanced output still under load, results for the unbalanced output were the same, except the gradual peak reached +4 dB at 11.7 KHz and high frequency response was extended, crossing -1 dB at 24.7 KHz. Note that the low-frequency response measured poorer than expected; I did not have time to trace the stage(s) responsible for this before shipping the modified unit. (None of the new coupling capacitors should be to blame.) Nor did I formally investigate the tone controls' effect on frequency response (the 'scope test with a square wave is a "quick and dirty" method). Probably, a gentle clockwise twist of the Bass knob will help compensate for the measured low-frequency "loss" (perhaps use its 12-o'clock position instead of the marked 11-ish position). As always, one's ears must be the final arbiter of what sounds best.

5.19. Channel 5: Noise and Distortion. Compared to CH1-CH4, CH5 is subjectively and objectively noisier. Since all five channels have a very similar triode gain stage at the front end, the main difference is due to the output driver stages, which are tube-based in CH5 (V3A and V4) and also contribute more voltage gain than the solid-state drivers of CH1-CH4. Additionally, any signal loss in the tone control network (which I did not attempt to measure) effectively degrades the channel's signal/noise ratio (SNR).

I measured noise under the conditions that yield 54.5 dB overall gain between CH5's input and balanced output as reported in Section 5.18 (i.e., feedback control set at the design level; balanced output set for nominal 600 Ω and terminated at 600 Ω), except the input was grounded instead of receiving a signal. Noise at the balanced output, as read on the Hewlett-Packard 331A's average-responding voltmeter, was corrected to represent true RMS (see Section 5.14; subtracting meter self-noise was not necessary given this channel's relatively high noise output). Noise was 2.03 mV_{RMS} (-53.8 dBV) with fader full clockwise and 1.01 mV_{RMS} (-59.9 dBV) with fader full counter-clockwise. An assumption that all measured noise is in the audio band seems reasonable, given the channel's frequency response described in Section 5.18. Assuming that the pre- and post-fader gain stages generate uncorrelated noise, the first triode stage's contribution, as amplified by the subsequent stages, is the square root of the difference between the two squared RMS readings given above, or 1.76 mV_{RMS}. That noise decreases with fader attenuation (turning fader counter-clockwise), while the relative effect of the post-fader-stages' constant noise (1.01 mV_{RMS}) increases. This makes the equivalent input noise (EIN) increase as the fader is turned down, as given in the following table:

Fader Attenuation, dB	EIN, dBV[1]	SNR, dB[2]
0 (full CW)	-108.3	74.3
-10	-103.3	69.3
-20	-94.3	60.3
-30	-84.4	50.4
-40	-74.4	40.4

1. Channel gain = 54.5 dB (design feedback level used; nominal 600-Ω balanced output terminated with 600 Ω).
2. With input amplitude giving 20.5 dBV balanced output into 600 Ω at 0dB fader attenuation, for easy comparison to CH1-CH4's noise figures given in Section 5.14. However, CH5's harmonic distortion at this output level is much higher than that of CH1-CH4 (see text).

CH5's -108.3 dBV EIN figure for the full clockwise fader is very close to the corresponding figure for the unbalanced input of CH1 or CH2 (-106.7 dBV; see Section 5.14). This reflects the similarity of the front-end triode stages and the full fader minimizing the output-stage's relative noise contribution. A more practical difference in CH5's noise compared to CH1/CH2 becomes apparent with fader attenuation applied. In addition, if one compares the signal-to-noise ratios (SNRs) of the channels for a given output amplitude, CH5's higher driver-stage gain causes a lower SNR even at full fader. To illustrate, the above table shows SNR computed for a 20.5-dBV output, so it may be directly compared to CH1-CH4's SNR given in the table in Section 5.14. At full faders, CH5's 74.3-dB SNR is 12.4 dB worse than that of CH1-CH4, which is 86.7 dB. From there, the SNR difference increases even more as faders are turned down, due to CH5's much noisier output driver as mentioned previously. Finally, a 20.5-dBV output from CH5 is much more distorted (on order of 3 %THD) than that of CH1-CH4 (near 0.5 %THD, see Figure 25). I will discuss CH5's distortion characteristics in general terms next.

Naturally, SNR improves as input signal amplitude increases (at any given fader setting). Of course, the trade-off is that distortion also increases. Given CH5's multi-triode signal path, the relationship between input level, fader setting, feedback setting, and output distortion is

complex. Unlike with CH1-CH4, I did not examine distortion in CH5 systematically, but can offer a few general comments and sporadic observations. The immediate post-fader triode stage (V3B) has limited headroom and is easily saturated; notice in Figure 14 that its cathode bias is only 83 mV. This is related to the triode's saturation threshold for signal peaks at the grid, although the amount of negative feedback applied to the cathode dynamically affects the "bias" (more feedback effectively increases the saturation threshold). In any case, input headroom at the V3B stage is not nearly as high as it is for the solid-state fader followers used in CH1-CH4; but in exchange, driver-stage clipping in CH5 is soft rather than hard.

Thus, CH5 offers multiple (but interacting) ways to tailor distortion effects: to emphasize input-stage (V3A) distortion, make the channel's input signal relatively high and set the fader relatively low. To emphasize output driver distortion, do the reverse; the amount of output-stage distortion is affected by both the feedback control (see Section 5.17) and the fader setting. In general, for any given output level, CH5 delivers more distortion than CH1-CH4. At the design feedback level and with balanced output set for the nominal 600 Ω and terminated at 600 Ω, distortion at 1 KHz measured: 0.45 % for a 8-dBV output using a -42.6-dBV input and arbitrary fader setting; it was 2.15 % for a 18.8-dBV output using a -35.5-dBV input and full fader setting.

5.20. Channel 5: Output Impedances and Transformer Characteristics. The low-Z balanced (XLR male) and "high-Z" unbalanced (1/4-inch) outputs of CH5 may be used separately or simultaneously. A dummy load resistor need not be hooked to the balanced output when using only the unbalanced output. The two outputs are in parallel, with the line transformer interceding for the balanced output. The output impedance of the final triode stage (V4, the 6CG7) thus determines the impedance of each output (reflected through the transformer in the case of the balanced output). I evaluated CH5's output impedances using three approaches: (1) The "direct" measurement method (as given in Section 5.9, second paragraph) for the balanced output at each setting of the output Z switch; (2) Inferring V4-stage output impedance from measured voltages across each side of the output transformer while driving a load at each output Z setting; and (3) calculating output impedance from published 6CG7 tube characteristics.

The "direct" Z_{OUT} measurements for the balanced output were taken at 1 KHz, with the feedback knob at the design level. With the output Z switch at the nominal 600 Ω setting (transformer secondary windings connected in series), Z_{OUT} measured 169 Ω. Set for the nominal 150 Ω output (the parallel connection), Z_{OUT} was 42 Ω. While these results are reassuringly close to an expected four-fold impedance difference between the two settings, the measured impedances are each nearly 3.6 times lower than their nominal values. This somewhat contradicts my assertion in Section 5.2 that vintage-era engineers aimed for maximum power transfer by matching source and load impedances; apparently, this was not exactly the case at the Altec 1567A's transformer-coupled output. There is certainly nothing wrong with driving a load from a 3.6-fold lower impedance.

As I will explain shortly, the second approach to determining output impedances also yields the transformer's voltage step-down ratio (which equals the "turns ratio," the number of turns in the primary winding per each effective turn in the secondary). Let me first report my DC resistance measurements on the transformer, because this affects the math. It also suggests that this

particular Altec type 15095 line output transformer may be defective or damaged slightly, even though it still works. I measured DC resistances by looking at voltage drops across windings supplied with actual DC, not by direct ohmmeter readings on my digital multi-meter; the latter uses pulses, giving inaccurate results for inductive devices. The DC resistances measured 1610 Ω for the primary, 35.0 Ω for the secondary between pins 1 and 3, and 47.9 Ω for the secondary between pins 4 and 6. Thus, an expected equal resistance for the two secondary windings was not observed; possible explanations are insulation failure (a short) between turns within the pins-1/3 secondary, or a high-resistance connection linking pins 4/6 to their winding. I did not have a pristine type 15095 unit for comparison.

In the second impedance-determining approach, I fed a 1-KHz sine wave to CH5's input, and with the feedback control at about the "1:30" position (flattest frequency response, slightly clockwise from the design level; see Section 5.17), hooked a 600-Ω load to the balanced output and measured the voltage at each output when the Z-out switch was in each position (series [nominal 600 Ω] versus parallel [nominal 150 Ω] secondary). With channel input amplitude and fader position constant but arbitrary, respective RMS voltages at the unbalanced and balanced outputs (i.e., primary and secondary sides of transformer) were: 24.25 and 4.00 for the series connection, and 26.0 and 2.48 for the parallel connection. Calculating impedances and turns ratios from this data requires four reasonable assumptions: (1) Switching from the series to parallel setting doubles the effective primary-to-secondary turns ratio. The nominal and observed four-fold difference in impedance between these settings (discussed above) supports this assumption, because reflected impedance is a function of the square of the turns ratio in an ideal transformer. (2) The output impedance of the final triode stage (V4, the 6CG7) and its source voltage were constant during the test. (3) The unbalanced output measurements represent the combined voltage drop across the 1610-Ω DC resistance of the primary (see above paragraph) in series with the balanced output's load impedance reflected through the transformer to the primary; with the load held constant, the latter is four-fold greater for the parallel versus the series case, consistent with the first assumption. (4) The effective load on the secondary is the load resistor (600 Ω for this test) in series with the DC resistance of the secondary, which is 82.9 Ω or 20.2 Ω for the series or parallel hookups, respectively, as suggested by the data in the preceding paragraph.

The data and assumptions given above are sufficient to simultaneously calculate the following: (1) A source (V4 stage output) impedance of 2.0 KΩ; and (2) reflected load impedances of 18.2 KΩ and 72.8 KΩ for the series and parallel secondary connections, respectively, when driving a constant 600-Ω load. Voltage drops across these impedances compared with those in the secondary (after accounting for voltage drops in the DC resistances) yield primary-to-secondary turns ratios of 4.90:1 and 9.94:1 for the series and parallel cases, respectively. This is quite close to 5:1 and 10:1 ratios predicted from the nominal impedances printed on the Altec type 15095 transformer and given in the original schematic (see Appendix), which are: primary, 15 KΩ; secondary, 600 Ω and 150 Ω for the series and parallel cases, respectively. If a 600-Ω load is supposed to reflect 15 KΩ on the primary as this suggests, the reason I obtained 18.2 KΩ could be due either to a slightly defective transformer (as suggested by the DC resistance readings discussed above), and/or errors in my measurements, which may be intensified by the exponential (square) when relating turns to impedances.

In any case, by calling the final triode stage's output impedance 2.0 KΩ, this second approach predicts balanced output impedances within 15 % of the ones directly observed in the first approach. Under the test conditions (feedback knob at "1:30," balanced output terminated with 600 Ω), CH5's unbalanced output Z at 1 KHz is 2.0 KΩ in parallel with 18.2 KΩ (assuming series connection), or 1.8 KΩ. It is simply 2.0 KΩ when the balanced output is unloaded. Importantly, however, all of these output impedances are affected by the feedback knob setting, as discussed next in connection with my third approach to evaluating CH5's output impedances.

The third approach involves modeling the performance of the 6CG7 (V4) stage based on this tube's published characteristics (I used General Electric spec sheet ET-T941B dated November, 1956). As shown in Figure 14 and described in Section 4.10, both triodes of the 6CG7 are wired in parallel to behave as a single triode, effectively halving the impedance compared to a single triode. Since the published characteristics are for individual triodes, I drew a load line on the plate characteristic chart representing current through one triode, as limited by a 30-KΩ plate load resistor (double the parallel configuration's 15-KΩ value). Also assuming a doubled cathode bias resistor value, the predicted operating point correlated well with the expected voltages (at B+, plate, and cathode) for the parallel combination. The analysis yielded a dynamic plate resistance (r_P) of 8.19 KΩ for a single triode; r_P for the parallel combination should be one-half this value, or 4.095 KΩ. In turn, placing this in parallel with the 15-KΩ plate load resistor predicts the V4-stage's open-loop (i.e., no feedback) output impedance: 3.22 KΩ.

However, negative feedback effectively reduces this output impedance, because it "tries" to keep the output voltage constant should load conditions change. (Imagine suddenly decreasing the load impedance. The output voltage would tend to drop, but this decreases the negative feedback signal, making the gain increase to compensate.) Theoretically, each 6-dB decrease in gain due to voltage feedback halves the output impedance. In the case of CH5, the full clockwise feedback setting is minimum feedback (maximum gain), not zero feedback (not open-loop). I did not attempt to test or model the gain reduction of this setting versus open-loop. Thus I can't calibrate (map to knob position) precisely how gain reduction (feedback increase as the control is turned counter-clockwise) reduces the V4-stage's predicted 3.22-KΩ open-loop output impedance. However, the observed 2.0-KΩ impedance for the "1:30" setting, just clockwise of the design feedback level, is at least approximately consistent with the 3.22-KΩ open-loop prediction. Also, one can predict that the 11-dB feedback knob range (see Section 5.17) should cause CH5's unbalanced output impedance to range by something like three-fold, from near 1 KΩ at maximum feedback, to near 3 KΩ at minimum. The corresponding impedance range for the balanced output is closer to two-fold, due to loss in the transformer primary's DC resistance, which acts in series with the V4-stage source impedance.

5.21. Channel 5: VU Meter. One obvious (and retro cool-looking) feature unique to CH5 is the illuminated VU meter. With two knobs on the vintage panel associated with it (illumination control and range switch), it's somewhat surprising that the VU meter itself was an optional accessory for the Altec 1567A (which the user can "install in minutes without soldering," as the original manual says; see Appendix for link). Under load, I measured CH5's balanced output when the meter read 0 VU at each range setting, and the results are in the following table; observing the vintage tradition (see Section 5.1), output levels are expressed in dBm:

VU Range Multiplier Setting	Output Into 600 Ω at 0 VU Indication[1], dBm
0	-1.0
+4	3.0
+8	7.2
+12	11.2

1. Test signal: constant (non-dynamic) 1.0 KHz sine waveform. Balanced output Z setting: nominal 600 Ω (secondary windings connected in series).

This data show that the range multiplier switch's 4-dB steps are accurate to within 0.2 dB, or about 2.3 % in terms of V_{RMS}, which is better than I expected. (And meter linearity was fairly good, as -6 VU indications corresponded with outputs averaging -6.37 dB of those giving 0 VU readings; data not shown.) The absolute output amplitudes may suggest that a 0-VU meter reading was originally calibrated to indicate a 0 dBm balanced output on the range 0 setting, 4 dBm on the +4 setting, et cetera. If the modified unit's type 15095 transformer is compromised (see Section 5.20), perhaps that accounts for 0.8 to 1.0-dB lower outputs than expected, if the expectation is 0 VU = 0 dBm for the balanced output terminated at 600 Ω. (The meter bridges the transformer's primary, not its secondary, as discussed next.)

Regarding how the VU meter monitors the output of the 6GC7 circuit instead of the line transformer's secondary, the original Altec 1567A manual (see Appendix for link) stated:

> "The VU multiplier is connected directly to the amplifier output rather than to the line side of the output transformer so that the VU meter may be used even though the 15095 transformer is not used. Very little compromise is made in the resistive termination of the meter even though the range multiplier is of a simple type. In the most sensitive position ('0' VU) the meter termination is 3450 ohms (11½ % low) and in the least sensitive position, 4150 ohms (6.4 % high), maintaining suitable ballistic characteristics."

A little algebra confirms that these "high" and "low" departure percentages point to 3900 Ω as the target source impedance ("termination") for the meter. The classic VU meters are designed to bridge a 600-Ω line when hooked in series with a 3600-Ω resistor; the instrument's source impedance is thus 3900 Ω (the 3600-Ω resistor in series with a driver-load network impedance of 300 Ω). This matches the meter's own internal impedance of 3900 Ω, for maximum power coupling to the meter. The Altec 1567A designers aimed for this source impedance because it affects the meter's transient response (i.e., its "ballistics"), which is a critical aspect of a VU meter's dynamic accuracy; for the classic VU meter characteristics, see Chapter 26 ("VU Meters and Devices" by Glen Ballou) in "Handbook for Sound Engineers, Third Edition," Glen M. Ballou, Editor, 2002, Focal Press (part of this chapter is included in a book preview available at http://books.google.com/).

At first, I was excited to read the manual's VU meter termination values, because these could be used to deduce Altec's evaluation of the final triode stage's impedance in parallel with the load reflected on the line transformer (if present). This might help confirm my own figures given in Section 5.20. However, for resistances used in the VU meter network (R35-R39 in original schematic [see Appendix], called R21-R25 in Figure 14), there is no single 6CG7 output network impedance value that satisfies both "3450 ohms" termination for the 0-VU setting and

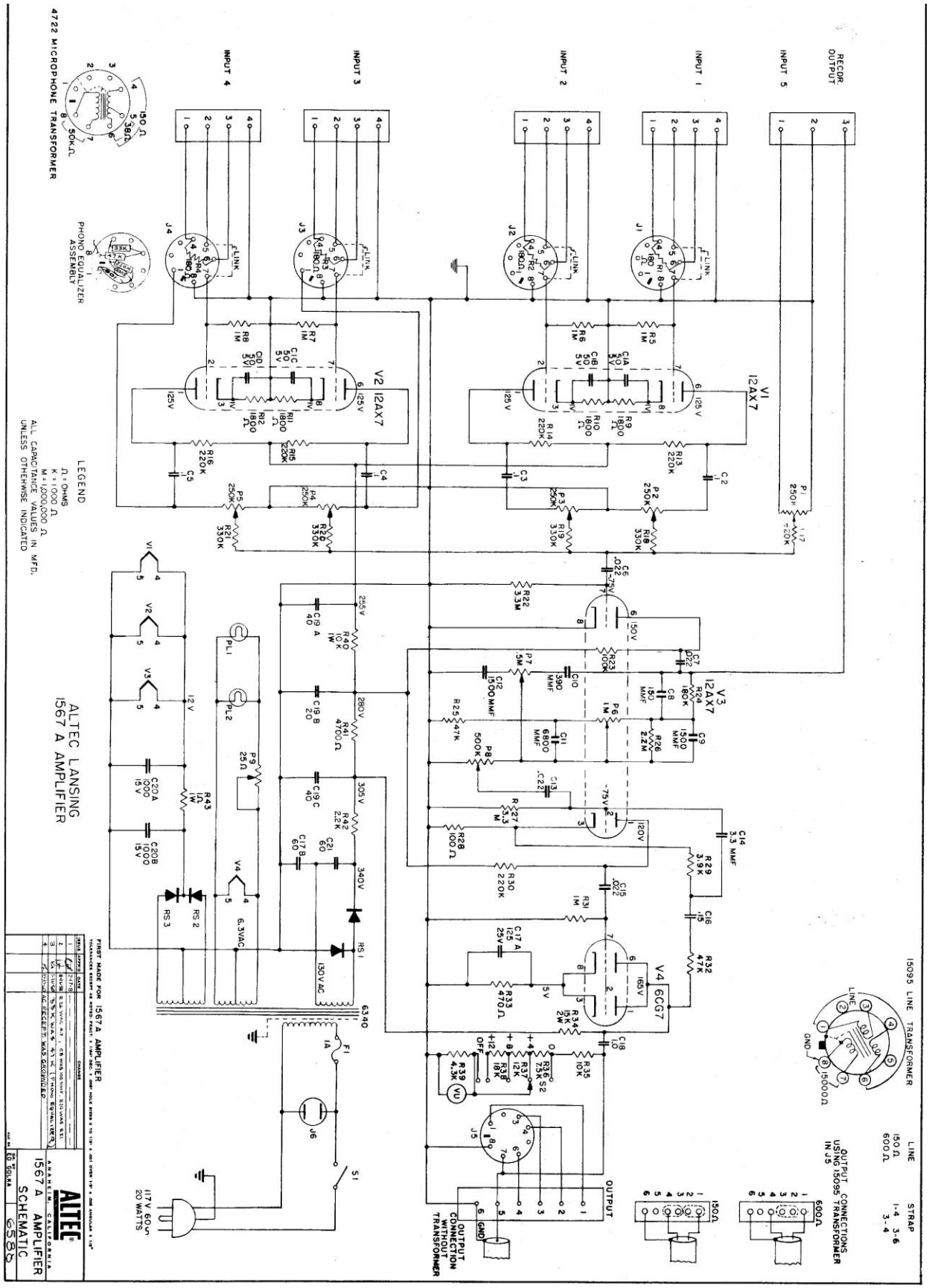
“4150 ohms” at +12 VU (the former predicts 7.45 KΩ and the latter a less likely 71.5 KΩ). Possibly, the original Altec 1567A manual made an error in one or both reported meter termination impedances. If the 6CG7 stage’s output impedance is 2.0 KΩ (see Section 5.20) and the load impedance reflected to the transformer’s primary is the nominal 15 KΩ, the output network impedance is 1.76 KΩ; in that case, VU meter termination is 3150 Ω (19.2 % under the target 3900 Ω) for the 0-VU setting, and 3950 Ω (1.3 % over target) for +12 VU.

6. Appendix: Original Altec 1567A Schematic

The original Altec 1567A manual and schematic is available online from AnalogRules.com:

http://www.analogrules.com/manuals/altec1.html

For your convenience, the original schematic is reproduced on the next page (page 62).


ALTEC LANSING
1567 A AMPLIFIER
SCHEMATIC
1567 A AMPLIFIER
ALTEC
ANAHEIM, CALIFORNIA
6586
4722 MICROPHONE TRANSFORMER
PHONO EQUALIZER ASSEMBLY
LEGEND
Ω = OHMS
K = 1,000 Ω
M = 1,000,000 Ω
ALL CAPACITANCE VALUES IN MFD. UNLESS OTHERWISE INDICATED
INPUT 1
INPUT 2
INPUT 3
INPUT 4
INPUT 5
RECDR. OUTPUT
V1 12AX7
V2 12AX7
V3 12AX7
V4 6CG7
6340
117 V 60~ 20 WATTS
OUTPUT CONNECTION WITHOUT TRANSFORMER
OUTPUT CONNECTIONS USING 1509S TRANSFORMER IN J5
150095 LINE TRANSFORMER
LINE
150 Ω
600 Ω
STRAP
1-4 3-6
3-4